95

(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

CORRECTED VERSION

(72) Inventors: GALARZA, Cecilia, Gabriela; Apt. 204, 39 Boardman Pl., San Francisco, CA 94103 (US). TSATSANIS, Michail; 444 Saratoga Ave. #35C, Santa Clara, CA 95050 (US). ERICKSON, Mark, Alan; 214 Linden Ave. San Bruno, CA 94066 (US). KANELLAKOPOULOS, Ioannis; 10152 Firwood Drive, Cupertino, CA 95014 (US). WAITE, James, W.; 402 Bella Vista Avenue, Los Gatos, CA 95032 (US). GU, Ming; 3200 Payne Ave. #729, San Jose, CA 95117 (US). SHAH, Sunil, C.; 637 Linden Avenue, Los Altos, CA 94022 (US). HERNANDEZ, Daniel, Joseph; 20425 Via Paviso #D12, Cupertino, CA 95014 (US). PARÉ, Thomas, E., Jr.; 229 Bush Street, Mountain View, CA 94041 (US). YUEN, Norman, Man, Leung; 10647-D Rosewood Road, Cupertino, CA 95014 (US). ROSARIO, Heberly; 926-B La Mesa Terrace, Sunnyvale, CA 94086 (US). LIN, Di; 4591 Pacific Rim Way, San Jose, CA 95121 (US). LOPEZ-DE-VICTORIA, Fernando; 4086 Farm Hill Blvd. #2, Redwood City, CA 94061 (US).

(74) Agent: MALLIE, Michael, J.; Blakely, Sokoloff, Taylor & Zafman LLP. 7th Floor. 12400 Wilshire Boulevard, Los Angeles, CA 90025 (US).

WO 01/35608 A1

(54) Title: METHOD AND APPARATUS FOR MITIGATION OF DISTURBERS IN COMMUNICATION SYSTEMS
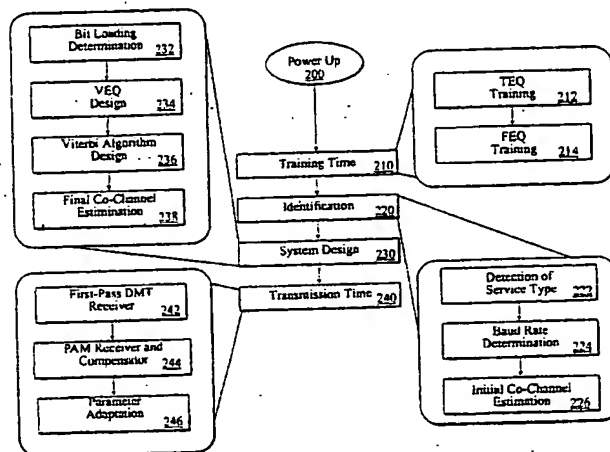
(57) Abstract: The present invention provides a system and method for compensating for cross-talk interference in communication systems. The compensation method consists of the following main phases: training time (210), identification (220), system design (230), and data transmission time (240). During training time (210), Time Domain Equalizer training (212) and Frequency Domain Equalizer training (214) are performed. Following the training time (210), the identification phase (220) encompasses the detection of existing disturbers and the estimation of their associated transmission parameters. The system design phase (230) entails the actual iterative design of several components of a compensator module. Lastly, the data transmission time phase (240) encompasses compensation of the Pulse Amplitude Modulated disturbance signals and final detection of the Discrete Multi Tone signal in the Pulse Amplitude Modulated compensated environment.

(81) Designated States *(national)*: AE. AG. AL. AM. AT. AU. AZ. BA. BB. BG. BR. BY. BZ. CA. CH. CN. CR. CU. CZ. DE. DK. DM. DZ. EE. ES. FI. GB. GD. GE. GH. GM. HR. HU. ID. IL. IN. IS. JP. KE. KG. KP. KR. KZ. LC. LK. LR. LS. LT. LU. LV. MA. MD. MG. MK. MN. MW. MX. MZ. NO. NZ. PL. PT. RO. RU. SD. SE. SG. SI. SK. SL. TJ. TM. TR. TT. TZ. UA. UG. UZ. VN. YU. ZA. ZW.

(84) Designated States *(regional)*: ARIPO patent (GH. GM. KE. LS. MW. MZ. SD. SL. SZ. TZ. UG. ZW). Eurasian patent (AM. AZ. BY. KG. KZ. MD. RU. TJ. TM). European patent (AT. BE. CH. CY. DE. DK. ES. FI. FR. GB. GR. IE. IT. LU. MC. NL. PT. SE. TR). OAPI patent (BF. BJ. CF. CG. CI. CM. GA. GN. GW. ML. MR. NE. SN. TD. TG).

Published:
— *with international search report*

(48) Date of publication of this corrected version:
30 May 2002

(15) Information about Correction:
see PCT Gazette No. 22/2002 of 30 May 2002. Section II

For two-letter codes and other abbreviations. refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

METHOD AND APPARATUS FOR MITIGATION OF DISTURBERS IN COMMUNICATION SYSTEMS

This application claims the benefit of the filing date of the following Provisional U.S. Patent Applications:

"IMPROVEMENTS IN EQUALIZATION AND DETECTION FOR SPLITTERLESS MODEM OPERATIONS", application number 60/165,244, filed November 11, 1999;

"CROSS-TALK REDUCTION IN MULTI-LINE DIGITAL COMMUNICATION SYSTEMS", application number 60/164,972, filed November 11, 1999;

"CROSS-TALK REDUCTION IN MULTI-LINE DIGITAL COMMUNICATION SYSTEMS", application number 60/170,005, filed December 9, 1999;

"USE OF UNCERTAINTY IN PHYSICAL LAYER SIGNAL PROCESSING IN COMMUNICATIONS", application number 60/165,399, filed November 11, 1999;

"CROSS-TALK REDUCTION AND COMPENSATION", application number 60/186,701, filed March 3, 2000;

"BLIND METHOD OF ONLINE COMPENSATION FOR INTERNAL CABLE BINDER IMPAIRMENTS", application number 60/215,514, filed June 30, 2000;

"FREQUENCY DOMAIN CROSS-TALK COMPENSATION METHOD", application number 60/215,633, filed June 30, 2000; and

"TIME DOMAIN CROSS-TALK COMPENSATION METHOD", application number 60/215,637, filed on June 30, 2000.

BACKGROUND OF THE INVENTION

1.   Field of the Invention

The present invention pertains to the field of communication systems. More

specifically, the present invention relates to a method and apparatus for mitigating disturbers

in communication systems.

2.   Background Information

The speed at which data is transmitted or received in digital communication systems

is significantly impaired by the level of background noise, impulse noise, cross-talk

interference, ingress noise coming from appliances, AM radio, and other communication

devices.

For example, in Digital Subscriber Line (DSL) configurations, cross-talk interference

arises from electromagnetic coupling of physically proximate channels. In DSL systems, a

data signal running along a telephone wire may be diminished by the noise that is injected by

the other signals running on adjacent wires. The cross-coupling between two channels can

create a highly correlated noise source that can degrade the performance of the transceiver

and, in severe instances, completely disable the main communication channel. Cross-talk

interference degrades the signal-to-noise ratio (SNR) of a data signal. Cross-talk interference

may also shorten the distance the signal can be received reliably, i.e., it may limit loop reach.

Additionally, cross-talk interference limits the bit rate for a given maximum allowable

transmit power. Such limitations may lower the number of users for a particular system and may limit the deployment of communication systems in certain regions.

When analyzing a single line in a network system, the cross-talk from adjacent lines is considered a disturber or noise. If a modem operating on an impaired line has access to the disturber, it may be able to cancel the interference through adaptive filtering techniques. Such measurements, however, are not always possible due to the lack of physical proximity of modems within a network.

Other factors that influence the data transmission through a network are: the large number of network users, the large amount of data collected from the deployed lines, and the presence of competing providers in the same physical line plant. The coexistence of ILECs (Incumbent Local Exchange Carriers) and CLECs (Competitive Local Exchange Carriers) in the same cable binders, brought about by the federally mandated deregulation of local telecommunication markets, implies that services deployed by one carrier may be disturbing the users of another carrier, who has no information about the source of this disturbance.

Thus, it is highly desirable to sort through the collected data and determine whether a specific line is being disturbed by external interference and whether that offending service belongs to the same carrier or not. Specifically, it is necessary to diagnose where the cross-talk problems are originating, predict how cross-talk affects services, and optimize the use of the system based on these predictions.

SUMMARY OF THE INVENTION

The present invention includes a method and system for compensating for cross-talk interference in communication systems. The method includes determining an estimation of at least one interfering signal and performing a compensation operation on the at least one interfering signal.

BRIEF DESCRIPTION OF THE DRAWINGS

The present invention is illustrated by way of example and not limitation in the accompanying figures in which:

Figure 1 illustrates a simplified diagram of an exemplary communication network;

Figure 2 illustrates a flow diagram of an interference compensation method according to one embodiment of the present invention;

Figure 3 illustrates a block diagram of one embodiment of a system implementing the compensation method illustrated in Figure 2;

Figure 4 illustrates a flow diagram of a method of generating a transmitted signal according to another embodiment of the present invention;

Figure 5 illustrates a flow diagram of an exemplary transmission path of the received signal;

Figure 6 illustrates a block diagram of a compensation architecture according to one embodiment of the present invention;

Figure 7 illustrates a flow diagram of a successive signal cancellation scheme according to one embodiment of the present invention;

Figure 8 illustrates a flow diagram of a method for determining a compensated signal-to-noise ratio according to yet another embodiment of the present invention;

Figure 9 illustrates a flow diagram of a bit loading method according to yet another embodiment of the present invention;

Figure 10 illustrates a flow diagram of an aggregation method according to the concepts of the present invention;

Figure 11 illustrates a block diagram of a joint viterbi design according to yet another embodiment of the present invention;

Figure 12 illustrates a flow diagram of a viterbi equalizer design procedure according to the concepts of the present invention;

Figure 13 illustrates a successive cancellation architecture of disturber signals of one embodiment of the present invention;

Figure 14 illustrates a MMSE VEQ design scheme for detecting multiple disturbers, according to yet another embodiment of the present invention;

Figure 15 illustrates another embodiment of the present invention where the compensation method is applied to a system with uncertainty;

Figure 16 illustrates a single user DFE design according to the concepts of the present invention;

Figure 17 illustrates a joint MIMO DFE design according to the concepts of the present invention;

Figure 18 illustrates a first pass DMT removal procedure according to the concepts of the present invention;

Figure 19 illustrates an embodiment of a possible architecture of a disturber remodulation and removal module;

Figure 20 illustrates a block diagram of a direct adaptation method according to the concepts of the present invention;

Figure 21 illustrates a block diagram of a indirect adaptation method according to the concepts of the present invention;

Figure 22 illustrates an exemplary embodiment of a direct adaptation mechanism according to the concepts of the present invention;

Figure 23 illustrates an exemplary communication system; and

Figure 24 illustrates an exemplary embodiment of the present invention as implemented in a DSL system.

DETAILED DESCRIPTION

A method and system for mitigation of disturbers in communication systems is disclosed. In the following description, for purposes of explanation, numerous specific details are set forth in order to provide a thorough understanding of the present invention. It will be evident, however, to one skilled in the art that the present invention may be practiced without these specific details. In some instances, well-known structures and devices are shown in block diagram form, rather than in detail, in order to avoid obscuring the present invention. These embodiments are described in sufficient detail to enable those skilled in the art to practice the invention, and it is to be understood that other embodiments may be utilized and that logical, mechanical, electrical and other changes may be made without departing from the scope of the present invention.

Some portions of the detailed descriptions that follow are presented in terms of algorithms and symbolic representations of operations on data bits within a computer memory. These algorithmic descriptions and representations are the means used by those skilled in the data processing arts to most effectively convey the substance of their work to others skilled in the art. An algorithm is here, and generally, conceived to be a self-consistent sequence of acts leading to a desired result. The acts are those requiring physical manipulations of physical quantities. Usually, though not necessarily, these quantities take the form of electrical or magnetic signals capable of being stored, transferred, combined, compared, and otherwise manipulated. It has proven convenient at times, principally for

reasons of common usage, to refer to these signals as bits, values, elements, symbols, characters, terms, numbers, or the like.

It should be borne in mind, however, that all of these and similar terms are to be associated with the appropriate physical quantities and are merely convenient labels applied to these quantities. Unless specifically stated otherwise as apparent from the following discussion, it is appreciated that throughout the description, discussions utilizing terms such as "processing" or "computing" or "calculating" or "determining" or "displaying" or the like, refer to the action and processes of a computer system, or similar electronic computing device, that manipulates and transforms data represented as physical (electronic) quantities within the computer system's registers and memories into other data similarly represented as physical quantities within the computer system memories or registers or other such information storage, transmission or display devices.

The present invention can be implemented by an apparatus for performing the operations herein. This apparatus may be specially constructed for the required purposes, or it may comprise a general purpose digital signal processor computer, selectively activated or reconfigured by a computer program stored in the computer. Such a computer program may be stored in a computer readable storage medium, such as, but not limited to, any type of disk including floppy disks, optical disks, CD-ROMs, and magnetic-optical disks, read-only memories (ROMs), random access memories (RAMs), EPROMs, EEPROMs, magnetic or optical cards, or any type of media suitable for storing electronic instructions, and each coupled to a computer system bus.

The algorithms and displays presented herein are not inherently related to any particular computer or other apparatus. Various general purpose systems may be used with programs in accordance with the teachings herein, or it may prove convenient to construct more specialized apparatus to perform the required method. For example, any of the methods according to the present invention can be implemented in hard-wired circuitry, by programming a general purpose processor or by any combination of hardware and software. One of skill in the art will immediately appreciate that the invention can be practiced with computer system configurations other than those described below, including hand-held devices, multiprocessor systems, FPGAs or other hardware platforms, microprocessor-based or programmable consumer electronics, network PCs, minicomputers, mainframe computers, and the like. The invention can also be practiced in distributed computing environments where tasks are performed by remote processing devices that are linked through a communications network. The required structure for a variety of these systems will appear from the description below.

The methods of the invention may be implemented using computer software. If written in a programming language conforming to a recognized standard, sequences of instructions designed to implement the methods can be compiled for execution on a variety of hardware platforms and for interface to a variety of operating systems. In addition, the present invention is not described with reference to any particular programming language. It will be appreciated that a variety of programming languages may be used to implement the teachings of the invention as described herein. Furthermore, it is common in the art to speak of software, in one form or another (e.g., program, procedure, application...), as taking an

action or causing a result. Such expressions are merely a shorthand way of saying that execution of the software by a computer causes the processor of the computer to perform an action or produce a result.

Overview of General Communication Network

The present invention is applicable to a variety of communication systems, for example: wireline, wireless, cable, and optical. Figure 23 illustrates an exemplary communication system 2305 that may benefit from the present invention. The backbone network 2320 is generally accessed by a user through a multitude of access multiplexers 2330 such as: base stations, DSLAMs (DSL Access Mulitplexers), or switchboards. The access multiplexers 2330 communicate management data with a Network Access Management System (NAMS) 2310. The NAMS 2310 includes several management agents 2315 which are responsible for monitoring traffic patterns, transmission lines status, etc. Further, the access multiplexers 2330 communicate with the network users. The user equipment 2340 exchanges user information, such as user data and management data, with the access multiplexer 2330 in a downstream and upstream fashion. The upstream data transmission is initiated at the user equipment 2340 such that the user data is transmitted from the user equipment 40 to the access multiplexer 2330. Conversely, the downstream data is transmitted from the access multiplexer 2330 to the user equipment 2340. User equipment 2340 may consist of various types of receivers that contain modems such as: cable modems, DSL modems, and wireless modems.

The invention described herein provides a method and system for managing the upstream and downstream data in a communication system. As such, the present invention provides management agents that may be implemented in the NAMS 2310, the access multiplexers 2330, and/or the user equipment 2340. One example of such a management agent is a system software module 2370 that may be embedded in the NAMS 2310. Another management agent that manages the data in the communication system 2305 is a transceiver software module 2360 that may be embedded in the access multiplexer 2330 and/or the user equipment 2340. Further details of the operation of modules 2370 and 2360 are described below.

For illustration purposes and in order not to obscure the present invention, an example of a communication system that may implement the present invention is a DSL communication system. As such, the following discussion, including Figure 23 and Figure 24, is useful to provide a general overview of the present invention and how the invention interacts with the architecture of the DSL system.

Overview of DSL Example

The present invention may be implemented in software modules or hardware that DSL equipment manufacturers may then embed in their hardware. Thus, although Figure 24 illustrates the present invention as software, the present invention should not be limited thereto. It should also be noted that this patent application may only describe a portion or portions of the entire inventive system and that other portions are described in co-pending patent applications filed on even date herewith.

Figure 24 illustrates an exemplary embodiment of the present invention as

implemented in a DSL system. The DSL system consists of a network of components

starting from the Network Management System (NMS) 2410 all the way down to the

Customer Premise Equipment (CPE) 2450. The following is a brief description of how these

components are interconnected.

The Network Management System (NMS) 2410 is a very high level component that

monitors and controls various aspects of the DSL system through an Element Management

System (EMS) 2420. The NMS 2410 may be connected to several Central Offices (CO)

2430 through any number of EMSs 2420. The EMS 2420 effectively distributes the control

information from the NMS 2410 to the DSL Access Multiplexers (DSLAMs) 2433 and

forwards to the NMS 2410 network performance or network status indicia from the DSLAMs

2433. DSLAMs 2433 reside in a Central Office (CO) 2430, usually of a telecommunications

company. Alternatively, DSLAMs 2433 may reside in remote enclosures called Digital Loop

Carriers (DLC). The CO 2430 may have tens or hundreds of DSLAMs 2433 and control

modules (CM) 2432. A DSLAM 3033 operates as a distributor of DSL service and includes

line cards 2435 and 2436 that contain CO modems. The CO modems are connected to at

least one line 2445, but more frequently it contains several line cards 2435 and 2436 that are

connected to several lines 2445. Usually the lines 2445 are traditional phone lines that

consist of twisted wire pairs and there may be multiple lines 2445 in a binder 2440 and

multiple binders in a cable. The transmission cables act as packaging and protection for the

lines 2445 until the lines 2445 reach the Customer Premise Equipment (CPE) 2450. It should

be noted that a DSLAM 2435 does not necessarily have to be connected to lines 2445 in a

-14-

single binder 2440 and may be connected to lines in multiple binders 2440. The lines 2445 terminate at the CPE 2450 in transceivers that include CPE modems. The CPE 2450 may be part of or connected to residential equipment, for example a personal computer, and/or business equipment, for example a computer system network.

As discussed in the background section, communications systems often suffer from interference and/or impairments such as crosstalk, AM radio, power ingress noise, thermal variations, and/or other "noise" disturbers. The present invention or portions of the present invention provide the user the capability to analyze, diagnose and/or compensate for these interferences and/or impairments. It also provides the ability to predict and optimize performance of the communication system in the face of impairments.

As illustrated in Figure 24, the transceiver software of the present invention 2460, depending upon how implemented, may provide the user with the ability to analyze, diagnose, and compensate for the interference and/or impairment patterns that may affect their line.

Also as illustrated in Figure 24, the system software 2470, depending upon how implemented, may provide the service provider with the ability to diagnose, analyze, and compensate for the interference and/or impairment patterns that may affect the service they are providing on a particular line. The diagnosis and analysis of the transceiver software also provide the ability to monitor other transmission lines that are not connected to the DSLAMs or NMS but share the same binders.

It should be noted that the system software 2470 may be implemented in whole or in part on the NMS 2410 and/or EMS 2420 depending upon the preference of the particular

service provider. Likewise, it should be noted that the transceiver software of the present invention 2460 may be implemented in whole or in part on the DSLAM 2433 and/or transceivers of CPE 2450 depending upon the preference of the particular user. Thus, the particular implementation of the present invention may vary, and depending upon how implemented, may provide a variety of different benefits to the user and/or service provider.

It should also be noted that the system software 2470 and the transceiver software of the present invention 2460 may operate separately or may operate in conjunction with one another for improved benefits. As such, the transceiver software of the present invention 2460 may provide diagnostic assistance to the system software 2470. Additionally, the system software 2470 may provide compensation assistance to the transceiver software of the present invention 2460.

Thus, given the implementation of the present invention with respect to the DSL system example of Figure 24, one of ordinary skill in the communications art would understand how the present invention may also be implemented in other communications systems, for example: wireline, wireless, cable, optical, and other communication systems. Further details of the present invention are provided below. Additional examples of how the present invention may be implemented in a DSL system are also provided below for illustrative purposes.

The present invention includes a method and apparatus for improving the quality of a digital signal on a main transmission line of a communication system by identifying the external disturbance signal and applying an opposite signal to compensate for the effect of the interference.

It will be appreciated that the compensation method disclosed herein may be used in various digital communication systems, such as: DSL, wireless, wireline, optical, or cable systems. However, the applicability of the present invention is not limited to such systems. Figure 1 illustrates a typical DSL network that may benefit from the cross-talk compensation method disclosed herein. The DSL network illustrated in Figure 1 consists of a Central Office (CO) 110 that is responsible for the management of the DSL system and provides services to the Customer Premises (CPE) 120. The CPE 120 consists of modems which contain DSL transceivers 120 responsible for 2-way transmission between the CPE lines and CO 110. It should be noted that the compensation method disclosed herein may be used at the transceiver level in any chip set that is directly connected to the signal line. Thus, DSL transceivers 120 that receive a disturbed signal may compensate for the cross-talk interference by using the present invention.

For illustration purposes, the following description depicts an example of a main transmission channel that is impaired by one or more disturbers, chosen to be controlled by an Asymmetric DSL (ADSL) modem. An ADSL channel may be characterized as a Discrete Multi Tone (DMT) channel. Using the cross-talk compensation method disclosed herein, disturbers that may be attenuated include, but they are not limited to: T1, E1, ISDN, or other DSL lines. These disturbers may be characterized as Pulse Amplitude Modulated (PAM) signals. It will be appreciated that the present invention also applies to disturbers that employ other modulation schemes such as QAM (Quadrature Amplitude Modulation), CAP, etc.

Overview of the Compensation Method

-17-

Figure 2 illustrates a flow diagram of the compensation method according to one

embodiment of the present invention. The compensation method consists of the following

main steps: training time 210, identification 220, system design 230, and data transmission

time 240. The initial channel training time 210 is performed after the modem is powered up

at step 200, as part of normal transceiver operation. During training time 210, Time Domain

Equalizer (TEQ) training 212 and Frequency Domain Equalizer (FEQ) training 214 are

performed, i.e., adaptation of the TEQ and FEQ parameters. Training time 210 further

encompasses the estimation of the DMT signal and of the SNR of each frequency slot (bin)

of the DMT signal. Following the training time 210, the identification phase 220

encompasses the detection of existing disturbers and the estimation of their associated

transmission parameters. For example, during the identification phase 220, the active

disturbers are detected at step 222 and their baud-rate determination 224 is performed.

Additionally, an initial estimation of the co-channel impulse responses 226 is performed.

More detailed descriptions of the training phase 210 and identification phase 220 are

described in co-pending Patent Application Serial No.(___), filed on even date herewith,

entitled " ", assigned to the assignee herein.

The system design phase 230 of the compensation method entails the actual iterative

design of several components of a compensator module. The compensator module of the

present invention may be located in a transceiver or in access multiplexers as illustrated in

Figure 23 and Figure 24. Given the channel information from the Identification phase 220, a

bit loading determination is performed at step 232. The bit loading is determined in order to

achieve an acceptable first-pass DMT error rate and produce the desired bit rate or margin

-18-

improvement. At step 234, a Viterbi Equalizer (VEQ) filter is designed in order to shorten the co-channel length and improve the SNR of the disturber signal. Next, at step 236, a Viterbi computational method is performed to detect the PAM disturber symbols. The PAM symbols detection is necessary for a more accurate data-aided final estimation 238 of other co-channels.

Lastly, the data transmission time phase 240 encompasses compensation of the PAM disturbance and final detection of the DMT signal in the PAM compensated environment. At step 242, a first-pass DMT receiver operation is performed. At step 244, during the PAM receiver and compensator phase, the adaptive VEQ processing and compensation of the detected PAM signal are performed. At step 246, the parameter adaptation is performed. A more detailed description of the transmission time phase 240 is described later below.

Figure 3 illustrates a block diagram of a system implementing the compensation method illustrated in Figure 2. The received signal y(t) 310 is first processed in a standard receiver 300. The signal y(t) 310 is first passed through an AD converter 315 and then it is filtered through a time domain equalizer (TEQ) 320. The output of the filter 320 is later utilized by the compensation module 350. The received signal is further processed with a prefix strip and Fast Fourier Transforms (FFT) 330. The signal is then filtered through a frequency domain equalizer (FEQ) 340. It will be appreciated that the operation of a standard communication receiver is well known by a person of ordinary skill in the art and, thus, further details of a receiver operation will not be described herein.

The compensation module 350 receives the initial co-channel estimations from the Identification Module 360 and the processed received signal from the standard receiver 300

and estimates a sequence of the disturber symbols for each of the disturbers that are chosen to be compensated. Then, this estimate of the disturber signal is subtracted from the main signal and the resulting compensated signal is passed on to the QAM decoder 362. The operation of the compensation module 350 is illustrated in Figure 10, Figure 11, and Figure 12. Further, the compensation module 350 estimates a compensated SNR of the received signal which is necessary in determining the compensated bit loading performed by the bit loading module 370. The methods of determining the compensated SNR and the bit loading are later described with reference to Figure 8 and Figure 9. Finally, the signal is processed by a QAM (Quadrature Amplitude Modulation) decoder or slicer 362 in order to obtain the compensated main channel symbols q(n) 385.

It should be noted that Figure 3 is meant to be illustrative and not limiting of the present invention. As such, other configuration may be used and other systems exhibiting interference and/or impairment problems may also benefit form the use of the present invention.

Received Signal Model

The signal received at the transceiver consists of a large number of components originating from the original DMT signal, the interference PAM signals, and noise. The received continuous time signal y(t) is represented by equation (1):

$$y(t) = y_{dmt}(t) + y_{pam}(t) + v(t) \quad , \tag{1}$$

where $y_{dmt}(t)$ denotes the received DMT signal, $y_{pam}(t)$ denotes the received PAM signal and $v(t)$ denotes un-modeled noise.

1. DMT Signal Generation

Figure 4 illustrates a flow diagram of how the transmitted DMT signal is generated. It should be noted that the use of a DMT signal in conjunction with the present invention is only an example and the present invention may also be used with other types of signals. Block 410 represents the mapping process of the transmitted DMT signal $b(n)$ 405 from bits to Quadrature Amplitude Modulation (QAM) symbols. At block 420, the output signal 415 of block 410 is then converted from a serial to a parallel configuration. For example, in one embodiment, the vectors are created of length 512, equal to the length of the DMT symbol. This operation is performed by selecting 223 elements out of 255 elements of the signal 415, prefixing them with 32 zeros, and conjugate them symmetrically, thus, extending the vector to 512 elements as shown in equation (2):

$$q(n) = [0,...,0, q(223n), q(223n+1),...,q(223n+222), 0, q^*(223n+222),...,q^*(223n), 0,...0]$$

$$(2)$$

At blocks 430 and 440, the vector $q(n)$ 425 is then processed by a diagonal gain matrix G and an IFFT (Inverse Fast Fourier Transform) matrix F, respectively. The resulting signal $x(n)$ 445 is given by equation (3):

$$x(n) = FGq(n)$$, $$(3)$$

where the diagonal gain matrix G is given by equation (4):

-21-

$$G = diag\{ 0, ..., 0, g(0), ..., g(223), 0, g^*(223), ..., g^*(0), 0, ..., 0\} \qquad (4)$$

and the IFFT matrix F is given by equation (5):

$$F = \frac{1}{N} \{ e^{j\frac{2\pi}{N}km} \}_{k,m = 0, ..., N-1} \qquad (5)$$

where $N = 512$. Finally, at block 450, a prefix of 32 samples is added to signal $x(n)$ 445, resulting in the final transmitted vector $s(n)$ 455 given by equation (6):

$$s(n) = \Pi_1 x(n) \qquad , \qquad (6)$$

where

$$\Pi_1 = \begin{bmatrix} 0 & I_{32 \times 32} \\ I_{512 \times 512} \end{bmatrix}$$

where I denotes the identity matrix. Finally, vector $s(n)$ 455 is converted from parallel to serial resulting in the scalar transmitted signal $s(nN + k) = \{ s(n) \}_k$.

Figure 5 illustrates the transmission path of the signal $s(n)$ 455 through the modulating pulse 510, main channel response 520, and receiver filters such as analog front end filter 530. Thus, the received continuous time signal $y_{dmt}(t)$ is given by equation (7):

$$y_{dmt}(t) = \sum_{k = -\infty}^{\infty} s(k) h(t - kT_{dmt}) \qquad , \qquad (7)$$

-22-

where $h(t)$ is the combined filter impulse response of blocks 510, 520, and 530. In one embodiment of the present invention, $T_{dmt} = 1/(2.208 \times 10^6)$ sec. Finally, the received continuous time signal $y_{dmt}(t)$ is sampled through the A/D converter at block 540, resulting in the signal $y_{dmt}(n)$ 45 given by equation (8):

$$y_{dmt}(n) = y_{dmt}(t)|_{t=nT_{dmt}} = \sum_{k=-\infty}^{\infty} s(k) h(n-k) \| \quad , \tag{8}$$

where $h(n)$ is given by equation (9):

$$h(n) = h(t)|_{t=nT_{dmt}} \tag{9}$$

2. PAM Disturber Signal Model

The continuous time PAM disturber signal is given by equation (10):

$$y_{pam}(t) = \sum_{j=1}^{J} y_j(t) \quad , \tag{10}$$

where J denotes the number of disturbers that are explicitly modeled in the received signal.

Each individual PAM signal $y_j(t)$ is given by equation (11):

$$y_j(t) = \sum_{k=-\infty}^{\infty} s_j(k) h_j(t-kT_j) \quad , \tag{11}$$

where $s_j(k)$ denotes the transmitted PAM sequence of the j-th disturber through an overall co-channel response $h_j(t)$ and with symbol period T_j.

The sampled signal $y_{pam}(n)$ at the receiver is given by equation (12):

$$y_{pam}(n) = y_{pam}(t)\big|_{t = nT_{dmt}} = \sum_{j=1}^{J} y_j(n) \qquad (12)$$

where $y_j(n)$ is given by equation (13):

$$y_j(n) = \sum_{k=-\infty}^{\infty} s_j(k)\, h_j(nT_{dmt} - kT_j) \qquad (13)$$

In one embodiment of the present invention, the baud period Tj of each PAM disturber is assumed to be an integer multiple of the DMT baud period, i.e., $T_j = P_j T_{dmt}$, wherein Pj denotes an integer over-sampling factor. Hence the equation (13) becomes:

$$y_j(n) = \sum_{k=-\infty}^{\infty} s_j(k)\, h_j(n - kPj) \qquad (14)$$

where $h_j(n)$ is given by equation (15):

$$h_j(n) = h_j(t)\big|_{t = nT_{dmt}} \qquad (15)$$

It should be noted that if the baud period Tj of each PAM disturber is not an integer multiple of the DMT baud period, the signal is resampled at a multiple of the PAM disturber baud rate.

3. Noise Signal Generation

The sampled noise signal is given by equation (16):

$$v(n) = v(t)\big|_{t = nT_{dmt}/P} \qquad (16)$$

General Compensation Architecture

Figure 6 illustrates a block diagram of the compensation architecture according to one embodiment of the present invention. In one embodiment and for ease of explanation, the compensation architecture includes four design modules: the standard receiver module 600, the DMT removal module 660, the disturber symbol detection module 670, and the PAM remodulation and removal module 680. The discrete time signal y(n) 610 is first received at the standard receiver block 600. The received signal 610 is first passed through the TEQ 620 resulting in the filtered signal $y_{teq}(n)$ 642. At block 630, the signal is further processed with a prefix strip and Fast Fourier Transform (FFT). The signal is then filtered through the FEQ 640, resulting in signal $q_1(n)$ 644.

1. DMT Removal Module

Since, in this embodiment, the channel identification operations are performed after the TEQ and FEQ training, the DMT signal has to be removed in order to obtain the aggregated disturbance signal.

The purpose of the DMT removal is to extract as much of the DMT component from the received signal as possible, to produce a signal $\hat{y}_{dis}$ 665 which consists mainly of cross-talk disturbance. This cross-talk disturbance signal can then be used to detect the disturber symbols $\tilde{s}(k)$ 675.

The first step in the DMT removal is an initial (1st Pass) detection on the DMT symbols, as indicated by the slicer function 662. The 1st pass detection produces 256 symbols which are conjugated and modulated back into the time domain using an IFFT. It should be noted that, in one example, the IFFT results in a real vector of 512 points, 32

points less than a frame of the received vector $y(n)$ 610. This loss of information is due to the cyclic prefix stripping operation 630 that occurs in the standard receiver block 600. At block 664, after the remodulation operation, this cyclic prefix is added back to the vector, and the time vector then filtered through the equalized main channel $\hat{h}_{keq}$. The resulting signal 666 is then subtracted from the TEQ filtered signal $y_{TEQ}(n)$ 642 to form signal $\hat{y}_{dist}(n)$, the 1st pass estimation of the disturbance signal:

$$\hat{y}_{dist}(n) = y_{TEQ}(n) - \hat{y}_{DMT}(n) \qquad (17)$$

A more detailed description of a DMT removal procedure is described below in the "System Transmission Time Phase" portion of the present application and, in particular, in the section entitled "DMT Removal".

2. Disturber Symbol Detection Module

The output vector $\hat{y}_{dist}(n)$ 665 from the DMT Removal module 660 is then passed to the VEQ 672 and Joint Viterbi Algorithm 674, resulting in a vector estimate 675 of the disturber symbols $\bar{s}(n)$. The VEQ is an FIR filter that preconditions the vector $\hat{y}_{dist}(n)$ 665, shortens the co-channels, and improves the SNR of the disturber signal. The Joint Viterbi Algorithm 674 processes the received vector for multiple disturbance symbols simultaneously by using a search routine based on the maximum likelihood function.

In another embodiment of the present invention, the disturber symbols may be detected using a Multiple Input Multiple Output DFE (Decision Feedback Equalizer). This embodiment will be described below in the section entitled "Alternative to VEQ Design".

3. PAM Remodulation and Removal

The compensation vector is constructed by modulating the detected set of disturber

symbols $\tilde{s}(k)$ 675 through the co-channel models, converting this signal to the frequency

domain using FFT, and applying the FEQ scaling. The compensation vector is then

subtracted from the original vector $\hat{q}_1(n)$ 644, as illustrated in Figure 6. The resulting signal

$q_2(n)$ 685 represents the compensated received signal. A more detailed description of the

PAM remodulation and removal module is described below in the "System Transmission

Time Phase" portion of the present application and , in particular, the section entitled "PAM

Remodulation and Removal".

In a successive cancellation architecture of one embodiment of the present invention,

signal $q_2(n)$ 685 is passed through a second slicer (2nd pass) in order to get the compensated

DMT symbols. Figure 7 illustrates a successive cancellation compensator architecture. The

received, uncompensated signal $y(k)$ 700 is first detected by a DMT receiver 710 and then it

is remodulated by a DMT remodulator 720 (first pass detection). The resulting signal 713 is

subtracted from the signal 717, which represents the original signal $y(k)$ delayed by a delay

715. The resulting disturbance signal is then detected by a PAM receiver 730, remodulated

by a PAM remodulator 740, and subtracted from the delayed received signal 737. The

resulting DMT signal is again detected by a DMT receiver 750 and remodulated by a DMT

remodulator 760 (second pass detection). As such, using a successive cancellation scheme

the final DMT symbols 770 are detected. It will be appreciated that the successive

cancellation scheme disclosed herein may be used with both time domain remodulation and

frequency domain remodulation. Further, other cancellation schemes may be used to detect

the DMT symbols, such as: joint detection of the PAM and DMT symbols or frequency

domain subspace cancellation.

System Design Phase

1. Determination of Bit-Loading Parameters

The effectiveness of the compensation method disclosed herein is negatively affected

as the rate of the first-pass DMT errors increases. Thus, a certain bit loading rate needs to be

determined to yield an acceptable first-pass DMT error rate and, at the same time, produce

the desired bit rate or margin improvement. Thus, a method for computing the bit loading

and gains to be used in a cross-talk compensation architecture is disclosed.

The method disclosed herein satisfies both the first-pass and second-pass

requirements. Since the second-pass requirements depend on the compensated SNR, that

SNR has to be computed before the transceiver goes into data transmission operation.

However, in one embodiment of the present invention, compensation does not occur until

transmission time. Thus, in this embodiment, the compensated SNR must be predicted

before compensation takes place using the SNR of the PAM receiver as the reference point.

In order to determine the compensated SNR, the amount of energy of the PAM disturbers

that may be removed by the compensation method disclosed herein must be predicted.

Compensated SNR Computation

-28-

In one embodiment of the present invention, during the training phase of the compensation method, the uncompensated per-bin SNR S_1 at the output of the first-pass DMT receiver is first determined. Further, D_1^{tot}, the Power Spectral Density (PSD) of the total noise at the first-pass DMT receiver, and the PSD of each PAM disturbers to be compensated for, are also determined. The total PSD of the compensated and uncompensated disturbers is then determined using equations (18) and (19):

$$D_c^{PAM} = \sum_{compensated} D_i^{PAM} \tag{19}$$

$$D_r^{PAM} = \sum_{uncompensated} D_i^{PAM} \tag{18}$$

where D_c^{PAM} denotes the PSD of the PAM disturbers to be compensated for and D_r^{PAM} denotes the PSD of the remaining PAM disturbers (not compensated). The sum D^{PAM} of the PSDs of the PAM disturbers is then calculated using equation (20):

$$D^{PAM} = D_c^{PAM} + D_r^{PAM} \tag{20}$$

The PSD of the white noise D_n is thus given by equation (21):

$$D_n = D_1^{tot} - D^{PAM} \tag{21}$$

For a computed compensated SNR S_2, the corresponding compensated bit loading b_2 for each bin is given by equation (22):

wherein Γ denotes a user-defined parameter based on a desired bit error rate.

$$b_2 = \log_2\left(1 + \frac{S_2}{\Gamma}\right) \qquad (22)$$

This bit loading will result in first-pass DMT errors with a probability $P_e^{DMT_1}$ that is approximated by equation (23):

$$P_e^{DMT_1} = 1 - \left[1 - 2Q\left(\sqrt{\frac{3 S_1}{2^{b_2}-1}}\right)\right]^2 \qquad (23)$$

The PSD of the first-pass DMT errors $D_e^{DMT_1}$ is therefore given by equation (24):

$$D_e^{DMT_1} = P_e^{DMT_1} D^{DMT} = P_e^{DMT_1} S_1 D_1^{tot} \qquad (24)$$

Thus, the total noise at the input of the PAM receiver D_n^{PAM} is given by equation (25):

$$D_n^{PAM} = D_e^{DMT_1} + D_r^{PAM} + D_n \qquad (25)$$

The (scalar) SNR S_p of the PAM receiver is then computed as the ratio of the total power of the compensated PAM disturbers W_c^{PAM} over the total power of the noise at the input of the PAM receiver W_n^{PAM}, which is illustrated in equation (26):

$$S_p = \frac{W_c^{PAM}}{W_n^{PAM}}, \quad W_c^{PAM} = \int_{\omega_1}^{\omega_2} D_c^{PAM}(\omega)\,d\omega, \quad W_n^{PAM} = \int_{\omega_1}^{\omega_2} D_n^{PAM}(\omega)\,W(\omega)\,d\omega \qquad (26)$$

-30-

where the integration of the PSDs takes place over the frequency interval (ω1, ω2) where

the compensated PAM disturber PSD is nonzero, and $W(\omega)$ denotes a weighting function

that accounts for the fact that the PAM receiver has different noise sensitivity in different

parts of the PAM spectrum.

Utilizing the SNR determined in equation (26), the probability of error of the PAM

receiver (for 2B1Q disturbers) is approximated by the equation (27):

$$P_e^{PAM} = \frac{3}{2}Q\left(\sqrt{\frac{2}{5}}S_p\right) \qquad (27)$$

Using the probability of error found in equation (27), the PSD of the compensated PAM

disturber that is not removed by compensation by the PSD of the errors made by the

PAM receiver is determined by equation (28):

$$D_e^{PAM} = P_e^{PAM} D_c^{PAM} \qquad (28)$$

Finally, the corresponding compensated per-bin SNR S_2 is computed using equation

(29):

$$S_2 = \frac{D^{DMT}}{D_1^{tot} - D_c^{PAM} + D_e^{PAM}} = \frac{D^{DMT}}{D_n + D_r^{PAM} + D_e^{PAM}} \qquad (29)$$

It should be noted that the compensated SNR S_2 is a vector-valued function in the

case of DMT receivers. Therefore, in order to optimize the search over the values of S_2, a

vector-valued search method may be used. In one embodiment of the present invention, a

3. Bit Loading and Gain Selection

Compensated bit loading

After the disturber ranking procedure is completed, the highest-ranked disturbers are selected for compensation, i.e., the disturbers with the highest compensated SNR. The compensated SNR procedure illustrated in Figure 8 is repeated, this time using the disturbers that are selected for compensation. This produces a predicted compensated SNR S_2 given by equation (29) and a corresponding predicted compensated bit loading b2 given by equation (22).

Viterbi limit

In one embodiment of the present invention, the PAM receiver uses a Viterbi maximum-likelihood sequence estimator (MLSE). The compensated SNR method described above does not account for the fact that the performance of the Viterbi MLSE depends

$$b_v = \log_2\left(1 + \frac{S_1}{\Gamma_v}\right) \tag{30}$$

nonlinearly on the first-pass DMT errors and deteriorates significantly when the first-pass DMT bit error rate (BER) is higher than a specific given value. Therefore, a "Viterbi limit" on the compensated bit loading is determined to ensure that the BER is not exceeded. The Viterbi-limited bit loading b_v is determined using equation (30):

wherein the Γ_v denotes a threshold based on the desired first-pass DMT bit error rate.

Final bit loading

The final bit loading is determined by comparing the predicted compensated bit loading b_2 given by equation (22) and the Viterbi-limited bit loading given by equation (30) on a bin-by-bin basis, selecting the smallest of the two for each bin, and then rounding to the

$$b_f = \lfloor \min(b_2, b_V) + 0.5 \rfloor \qquad (31)$$

nearest bin. This can be expressed through equation (31):

where the floor brackets indicate the integer part of the argument, and the minimum operation is performed on a bin-by-bin basis.

Final gain selection

In a bit loading method for a DMT modem, the gain for each bin is chosen so that the resulting SNR will yield the desired BER for the selected bit constellation on this bin. Since the upper bound on the BER is expressed through the parameter Γ and the SNR before the gain is denotes as S, gain g that corresponds to the bit loading b is given by equation (32) :

$$g = \frac{\Gamma}{S}(2^b - 1) \qquad (32)$$

In one embodiment of the present invention, it is desirable to have the final gain g_f satisfy two conditions. As the first condition, the final gain has to guarantee that with the bit loading selected in equation (31), the BER of the first-pass receiver does not exceed the

$$g_f \geq g_{f1} = \frac{\Gamma_V}{S_1}(2^{b_f} - 1) \qquad (33)$$

Viterbi limit prescribed by Γ_v. Since the SNR at the first-pass DMT receiver is equal to the measured uncompensated SNR S_1, this condition is expressed by equation (33):

As the second condition, the final gain g_f has to guarantee that with the bit loading selected in (31), the BER of the second-pass receiver does not exceed the limit prescribed by Γ. Since the SNR at the second-pass DMT receiver is predicted to be the compensated SNR S_2, this condition is expressed by equation (34):

Thus, in this embodiment, the final gain is selected to satisfy both equations (33) and

$$g_f \geq g_{f2} = \frac{\Gamma}{S_2}(2^{b_f} - 1) \qquad\qquad (34)$$

(34) and the final gain chosen to be the maximum value of g_{f1} and g_{f2}. Thus, the final gain

$$g_f = \max (g_{f1}, g_{f2}) \qquad\qquad (35)$$

of this embodiment gf is given by equation (35):

Figure 9 is a flowchart of the compensated SNR and bit loading methods. Using the predetermined inputs 900, the disturbances are first ranked at step 910 in the order of how they affect the bit loading. At step 920, the highest ranked disturbers are selected for compensation and a predicted compensated bit loading b_2 942 is determined using the method illustrated in Figure 8. Further, using the uncompensated SNR and the SNR threshold 930, a Viterbi-limited bit loading b_v 944 is determined at step 940. The final bit loading is then determined at step 950 by comparing b_2 942 with b_v 944 on a bin-by-bin

basis and selecting the smallest of the two for each bin. Lastly, at step 960, the final gain g_f

is determined.

Disturber Receiver Design

1.Constellation Aggregation

In order to determine the estimates of the disturber symbols $s_i(n)$ of cross-talk

interference, the joint Viterbi algorithm (JVA) may be used. In one alternative

embodiment of the present invention, a MIMO (Multiple Input Multiple Output) DFE

(Decision Feedback Equalizer) method may be used as well. A MIMO DFE method is

described below with reference to Figures 16 and 17.

The computational load incurred by the JVA makes it difficult to implement when

more than two disturbers are present. In order to take advantage of the optimality of the JVA

while reducing its complexity, i.e., its number of states, constellation aggregation may be

performed. Since it is desirable to identify the combined effect of all the disturbers instead of

the contribution of each individual PAM disturber, it is more effective to find an

approximated signal constellation set with a much smaller size for all PAM disturbers by

minimizing the approximation error or distortion thus introduced. It should be noted that,

according to one embodiment of the present invention, constellation aggregation is

performed during the design phase of the compensation architecture.

The signal model $y(n)$ of PAM disturbers $s_i(n)$, $i = 1, \cdots, M$, is given by equation

(36):

$$y(n) = \sum_{i=1}^{M} \sum_{k=1}^{L} s_i(k) h_i(n-k), \qquad L = \max_i L_i. \qquad (36)$$

where M PAM disturbers symbols are all chosen from the same alphabet set of size A,

$h_i(n)$ denote the corresponding impulse responses of the M channels with order L_i, and

$y(n)$ is the summation of M overlapping pulses.

The received signal signatures $g_k(n)$ and $\underline{g}_k$ due to the transmitted PAM symbol at a

given time k, $s_i(k)$, are given by equations (37a) and (37b):

$$g_k(n) = \begin{bmatrix} s_1(k) & \cdots & s_M(k) \end{bmatrix} \begin{bmatrix} h_1(n) \\ \vdots \\ h_M(n) \end{bmatrix} \qquad (37a)$$

$$\underline{g}_k = \begin{bmatrix} g_k(0) & \cdots & g_k(L) \end{bmatrix}. \qquad (37b).$$

Because the vector $\begin{bmatrix} s_1(k) & \cdots & s_M(k) \end{bmatrix}$ has $N = A^M$ combinations, $\underline{g}_k$ can take one

of N values. With the definition of $g_k(n)$, equation (36) may be re-written as:

$$y(n) = \sum_{k=0}^{L} g_k(n-k). \qquad (38)$$

Next, the aggregated $\bar{g}_k(n)$ with $\underline{\bar{g}}_k = \begin{bmatrix} \bar{g}_k(0) & \cdots & \bar{g}_k(L) \end{bmatrix}$ taking N' values $(N' < N)$ is

determined such that the cost given by equation (39) is minimized.

$$C = \sum_n \left| y(n) - \bar{y}(n) \right|^2 \qquad (39)$$

-37-

Similar to $y(n)$, $\bar{y}(n)$ is defined by equation (40):

$$\bar{y}(n) = \sum_{k=0}^{L} \tilde{g}_k(n-k). \tag{40}$$

Expanding the cost function of equation (39) leads to

$$C = \sum_n \left| \sum_{k=0}^{L} [g_k(n-k) - \tilde{g}_k(n-k)] \right|^2 \tag{41}$$
$$\leq \alpha \sum_n \sum_{k=0}^{L} |g_k(n-k) - \tilde{g}_k(n-k)|^2,$$

where α is a scalar. Assuming that $g_k(n-k) \neq 0$ for $n = k, k+1, \cdots, k+L$, the above cost

may be expressed by equation (42):

$$C = \sum_{k=0}^{L} \left\| \underline{g}_k - \tilde{\underline{g}}_k \right\|^2. \tag{42}$$

It should be noted that $\underline{g}_k$ has N possibilities which can be denoted as a set

$G = \left\{ \underline{g}^{(j)} \right\}_{j=1}^{N}$, where each $\underline{g}^{(j)}$ in the set corresponds to one such possible value. Similarly,

there is an aggregated set $\tilde{G} = \left\{ \tilde{\underline{g}}^{(j)} \right\}_{j=1}^{N}$ and $\tilde{\underline{g}}_k \in \tilde{G}$. Thus, to minimize the original cost is

equivalent to minimize the following function given by equation (43):

$$C = \sum_{j=1}^{N} \left\| \underline{g}^{(j)} - \tilde{\underline{g}}^{(j_k)} \right\|^2, \tag{43}$$

-38-

where $\tilde{g}^{(j_k)}$ represents the corresponding aggregation from $\underline{g}^{(j)}$. Thus, the aggregation

method involves a clustering method which dictates that the optimal cluster centers $\tilde{\underline{g}}^{(k)}$ are

centroids. Specifically, a point in the original cluster $\underline{g}^{(j)}$ is given by equation (44):

$$\underline{g}^{(j)} = \begin{bmatrix} s_1^{(j)} & \cdots & s_M^{(j)} \end{bmatrix} \begin{bmatrix} \underline{h}_1 \\ \vdots \\ \underline{h}_M \end{bmatrix} \tag{44}$$

where $\underline{h}_i = \begin{bmatrix} h_i(0) & \cdots & h_i(L) \end{bmatrix}$ $(i = 1, \cdots, M)$.

Thus, the aggregated constellation point with $\tilde{s}_i(k) = \dfrac{1}{N_k} \sum_{j_k=1}^{N_k} s_i^{(j_k)}$ represents the

centroid equivalence given by equation (45):

$$\tilde{\underline{g}}^{(k)} = \frac{1}{N_k} \sum_{j_k=1}^{N_k} \underline{g}^{(j_k)} = \left(\frac{1}{N_k} \sum_{j_k=1}^{N_k} \begin{bmatrix} s_1^{(j_k)} & \cdots & s_M^{(j_k)} \end{bmatrix} \right) \begin{bmatrix} \underline{h}_1 \\ \vdots \\ \underline{h}_M \end{bmatrix} = \tilde{\underline{s}}(k) \begin{bmatrix} \underline{h}_1 \\ \vdots \\ \underline{h}_M \end{bmatrix}, \tag{45}$$

where $\tilde{\underline{s}}(k) = \begin{bmatrix} \tilde{s}_1(k) & \cdots & \tilde{s}_M(k) \end{bmatrix}$ and N_k is the number of points in the N-cluster

aggregating to $\tilde{\underline{g}}^{(k)}$ in the N'-cluster.

Given a set of (initial) centroids $\tilde{\underline{g}}^{(j)}$, $\underline{g}^{(j)}$ may be assigned to the centroid $\tilde{\underline{g}}^{(j_k)}$

such that the distance

$$d_k = \left\| \underline{g}^{(j)} - \tilde{\underline{g}}^{(k)} \right\|^2 \tag{46}$$

-39-

is minimized. Subsequently, based on the current cluster assignment, the centroids given by equation (47) may be recalculated:

$$\bar{\underline{s}}^{(k)} = \frac{1}{N_k} \sum_{j_k=1}^{N_k} \underline{s}^{(j_k)} .$$ (47)

It should be noted that the above cluster assignment and centroid recalculation can be processed iteratively until convergence. It will be appreciated that using the aggregation method disclosed herein, the size of constellation may be greatly reduced while the aggregation error is limited within an allowable range. For example, for five 4-PAM disturbers, the input signal can take one of $4^5 = 1024$ combinations. However, using the aggregation method disclosed herein, a reduced constellation set of 16 points may be obtained with only minimal power lost. It should be noted that, according to one alternative embodiment of the present invention, a K-means clustering method may be applied to perform the constellation aggregation.

Figure 10 is a flow diagram of the aggregation method. The inputs 1010 of the aggregation function are: the (original) constellation $C \in R^{N \times M}$ to be aggregated (C is formed from set S), the channel impulse response matrix, and the desired constellation size. During the aggregation process 1020, the aggregated constellation $\tilde{C} \in R^{N \times M}$ 1030 are generated from set $\bar{S}$ and the percentage of distortion introduced due to the aggregation process is determined.

2. VEQ design and Channel Shortening

As illustrated in Figure 6, the VEQ is an FIR filter that preconditions the vector $\hat{y}_{dist}(n)$ 665 of Figure 6. The VEQ method disclosed herein shortens the co-channels to a user-selected length so that the joint Viterbi algorithm (JVA) can operate efficiently with a minimum number of computational operations. Furthermore, the VEQ enhances performance of the joint Viterbi, thus increasing SNR at the input.

The JVA 674 processes the received vector for multiple disturbance symbols simultaneously by using a search routine based on the maximum likelihood sequence estimate (MLSE) for a symbol sequence. Typically, the JVA performance may be degraded by additive noise that does not have a flat spectrum, but has significant correlation, thus leading to symbol detection errors. Also, when the symbols are sent through a channel with a long impulse response (i.e., having many taps) the number of operations (multiplications, additions, etc.) required to implement the VA could become prohibitive. This constraint becomes even more problematic in the case of the JVA, when multiple communication paths are decoded simultaneously, since the number of operations grows exponentially with the number of channels. Thus, if either the white noise or short channel length assumption is not satisfied, effective use of the JVA may not be possible. In these cases, pre-filtering, or equalization of the JVA input signal may be necessary. Thus, a method to jointly optimize this filter (or equalizer) design and the JVA path metrics that can approximate optimality conditions and enable a pragmatic JVA implementation is disclosed herein.

Figure 11 illustrates a joint Viterbi algorithm. The symbols to be detected $d_1,...,d_m$ 1105 are passed through the corresponding channels $h_1,...,h_m$ 1115. Noise signal z

-41-

1120 is added to the output of these channels to produce the signal y 1145 at the receiver.

It should be noted that signal y 1145 may be oversampled, thus improving the effectiveness

of the VEQ design.

In general, the noise z 1120 may be comprised of several signal components.

When the Viterbi is used for cross-talk compensation for a DMT DSL modem, this noise

may consist of colored noise ($h_n * n$) 1135, other disturber channels

($h_{m+1} * d_{m+1} + \cdots + h_{m+p} * d_{m+p}$) 1130, and residual tones 1125 that result from first pass

DMT errors. These different sources of noise may be included in the VEQ design in the

manner described by equations (23)-(25).

It should be noted that the total noise, including noise signals 1130 and 1135 of

Figure 11, may be modeled by an aggregate source $\bar{n}$. The aggregate source $\bar{n}$ is passed

through a shaping filter $\bar{h}$ so that the spectrum of the received noise closely matches that

of a particular noise environment. The filter $\bar{h}$ is chosen to capture the aggregate spectrum

of the total noise signal z 1120.

The purpose of the JVA 1160 is to process the received signal y 1145 and decode

the sent symbols, thus producing estimated values $\hat{d}_1,...,\hat{d}_m$ 1170. However, when noise

shaping introduced by the channel $\bar{h}$ is significant, or when the co-channel models $h_1,...,h_m$

1115 have an excessive number of taps, proper functioning of the JVA can be

pragmatically prohibitive.

-42-

VEQ Design

Figure 12 illustrates a VEQ filter design procedure according to one embodiment of the present invention. Step 1210 is the initial set-up phase during which the design parameters are selected. The design parameters include: n_h, the desired channel lengths for co-channels, $n_{\bar{h}}$, the aggregate (correlated) noise channel, n_w, the VEQ filter length, the number of disturbers to compensate, and the first-pass DMT error rates and tone magnitudes.

Further, a relative weight $\beta \in [0,1]$ is selected to allow a design trade-off between noise whitening and channel shortening. For example, to achieve all noise whitening without shortening co-channels, the relative weight may be set at the value of 1. To shorten co-channels without regard to noise correlation, the relative weight may be selected as $\beta = 0$.

Next, channel convolution matrices $H_i, i = 1,...,m$, and noise correlation filter matrix, $\bar{H}$ are formed. It should be noted that these matrices can be formed in either row or column form. In one embodiment of the present invention, these matrices are coded with the channel impulse on the columns, as given by equation (48a):

$$H_i = \begin{bmatrix} h_1^i & 0 & & \\ h_2^i & h_1^i & & \\ \vdots & h_2^i & h_1^i & \\ 0 & \vdots & \vdots & h_1^i \end{bmatrix},$$

(48a)

where each matrix has dimension $R^{(n_{h^i}+n_w-1)\times n_w}$, where n_{h^i} is the length of the ith co-channel, and n_w is the specified length of the equalizer.

The co-channels may be normalized as illustrated in equation (48b), so that the effective shortening can be carried out evenly over the co-channels, and the relative weighting β is a more meaningful trade-off parameter.

$$h_i = \frac{h_i}{\|h_i\|_2}$$

(48b)

At step 1220, the viterbi equalizer w is solved using a joint least squares method as illustrated in equation (49). It should be noted that a singular value decomposition method may be used to solve for the equalizer w.

$$\begin{bmatrix} (1-\beta)\begin{bmatrix} H_1^r \\ \vdots \\ H_m^r \end{bmatrix} \\ \beta \overline{\overline{H}}^r \end{bmatrix} w = \begin{bmatrix} e_{\delta+1}^1 \\ \vdots \\ e_{\delta+1}^m \\ e_{\delta+1}^{m+1} \end{bmatrix}.$$

(49)

Matrices $H_i^r, i=1,...,m$, and $\overline{H}^r$ are the reduced matrices that are derived from channel convolution matrices $H_i, i=1,...,m$, and noise correlation matrix, $\overline{\overline{H}}$, by removing

-44-

$n_h - 1$ rows and $n_{\bar{h}} - 1$ rows, respectively, that correspond to the desired windows of co-channel response. This process is performed for a given pair of delays δ (for the co-channels) and $\bar{\delta}$ (for the composite noise channel). Equations (50a) and (50b) illustrate the matrix row removal process, according to one embodiment of the present invention.

$$H_i = \begin{bmatrix} H_i(1:\delta,:) \\ H_i(\delta+1:\delta+n_h,:) \\ H_i(\delta+n_h+1:end,:) \end{bmatrix}, \text{ and } \overline{H} = \begin{bmatrix} \overline{H}(1:\bar{\delta},:) \\ \overline{H}(\bar{\delta}+1:\bar{\delta}+n_{\bar{h}},:) \\ \overline{H}(\bar{\delta}+n_{\bar{h}}+1:end,:) \end{bmatrix} \tag{50a}$$

$$H_i' = \begin{bmatrix} H_i(1:\delta+1,:) \\ H_i(\delta+n_h+1:end,:) \end{bmatrix}, \text{ and } \overline{H}' = \begin{bmatrix} \overline{H}(1:\bar{\delta}+1,:) \\ \overline{H}(\bar{\delta}+n_{\bar{h}}+1:end,:) \end{bmatrix} \tag{50b}$$

for $i = 1,...,m$, by stripping out the middle matrices that consist of $n_h - 1$ and $n_{\bar{h}} - 1$ rows, respectively.

In addition, vectors $e_{\delta+1}^i$ are unit vectors of length equal to the number of rows in H_i', with the $\delta+1$ element, corresponding to the beginning of the desired window, set to unity in order normalize the equalized response. Unit vector $e_{\bar{\delta}+1}$ corresponds to the reduced matrix $\overline{H}'$ for the noise channel.

Next, at step 1260, the Least Squares Metric is determined, using equation (51), for any given equalizer w.

$$PERF_{LS} = \left\| H^{out} w \right\|_2 \Big/ \left\| H^{in} w \right\|_2 \tag{51}$$

-45-

where

$$H^{out} = \begin{bmatrix} H_1^{out} \\ \vdots \\ H_m^{out} \\ \overline{H}^{out} \end{bmatrix}, \text{ and } H^{in} = \begin{bmatrix} H_1^{in} \\ \vdots \\ H_m^{in} \\ \overline{H}^{in} \end{bmatrix} \qquad (52a)$$

with

$$H_i^{out} = \begin{bmatrix} H_i(1:\delta,:) \\ H_i(\delta + n_h + 1:end,:) \end{bmatrix}, \text{ and } \overline{H}^{out} = \begin{bmatrix} \overline{H}(1:\overline{\delta},:) \\ \overline{H}(\overline{\delta} + n_{\overline{h}} + 1:end,:) \end{bmatrix}, \qquad (52b)$$

and similarly,

$$H_i^{in} = H_i(\delta + 1:\delta + n_h,:), \text{ and } \overline{H}^{in} = \overline{H}(\overline{\delta} + 1:\overline{\delta} + n_{\overline{h}},:). \qquad (52c)$$

It should be noted that the delay values δ and $\overline{\delta}$ are swept over until $PERF_{LS}$ is either minimized, or achieves an acceptably low value.

The performance of the VEQ is then analyzed, at step 1260, by evaluating one or more of the following metrics: noise whiteness, equalized (shortened) co-channel length, or SNR at the input to the joint Viterbi processor.

At step 1280, the performance of the VEQ is evaluated. If the VEQ performance is not optimal or satisfactory, at step 1240, the VEQ filter length may be increased or the relative weight β may be varied. Satisfactory performance may be achieved, in one embodiment of the present invention, when repositioning the windows by incrementing δ

and $\bar{\delta}$, and repeating the process of determining the reduced matrices H_i', $i = 1,...,m$,

and $\overline{H}'$, and computing the VEQ w. If no window position meets desired performance, the

VEQ filter length n_w may be increased, and, as illustrated in Figure 12, steps 1220, 1260,

and 1280 are repeated. If channels are properly shortened, but the remaining correlation in

the equalized noise channel is unsatisfactory, the relative weight β may be increased and

steps 1220, 1260, and 1280 are repeated. Similarly, if channels are not shortened enough,

the relative weight β may be decreased and the design method may be re-executed.

The method described above performs joint equalization and noise whitening for

joint detection of multi-user communication channels. This method produces a linear

equalizer that performs the following functions: shortens the co-channels to a desired

length, thus reducing the JVA computational requirements, and improves detection

accuracy by whitening the additive noise. The method further reduces the effect of residual

DMT tones at the input of the Viterbi by attenuating those frequencies where there is high

probability of DMT error on a first pass decision, when the JVA is used as part of a cross-

talk compensation method. Moreover, this VEQ design method equalizes the co-channels

to the same window (in time) to improve SNR at the input and to simplify implementation

of the Viterbi detection method.

It should be noted that the VEQ design is set-up in matrix form and the optimal

VEQ filter is obtained by solving the least squares method (LSM). However, there are

alternate computational methods that may be implemented that utilize an MMSE

(Minimum Mean Square Error) formulation. In order to avoid the matrix inversions

-47-

associated with matrix operations that are potentially incompatible with fixed point processing, adaptive schemes, such as LMS, based on the MMSE may also be implemented.

An alternate method for an equalizer design for the Viterbi algorithm is thus further described with reference to Figure 14. Instead of the least squares method based on weighted channel matrices, the VEQ design is accomplished by minimizing the mean square error subject to a constraint on the length of the equalized channels. This approach can be applied to design equalizers for individual channels, or in a joint manner where a single filter shortens the impulse response of several channels simultaneously, or alternatively, in a successive manner for the different channels.

Figure 14 illustrates a VEQ design scheme for detecting a set of symbols for multiple disturbers, in a simultaneous manner, according to one embodiment of the present invention.   For each disturber s_1 and s_2, the target channel responses (or filters) b_1 1415 and b_2 1425 are designed in order to minimize the error in the VA 1440. The channels are equalized jointly with a single equalizer 1430. The target responses b_1 and b_2 and the VEQ parameters 1430 are designed simultaneously, such that the output of the VEQ filter is compared at block 1450 with the target responses b1 and b2. The MSE e(k) is evaluated in order to obtain a VEQ design that induces the least error in the VA 1440.

In order to find the VEQ filter parameters, the co-channel models h_1 1410 and h_2 1420 are determined. Signal y 1435, the input of the VEQ filter 1430, is thus formed by adding the co-channel models 1410 and 1420 with an additive gaussian noise n 1412.

The mean square error is defined as the expected value of the square of the error and it is given by equation (53):

$$MSE : Ee_k e_k^* \stackrel{\Delta}{=} \langle e_k, e_k \rangle \tag{53}$$

Given a set of target responses b_i, the optimal viterbi equalizer parameter w is given by equation (54):

$$w_{opt}^* = \begin{bmatrix} \tilde{b}_1^* & \cdots & \tilde{b}_m^* \end{bmatrix}_{opt} \begin{bmatrix} R_{s_1 y} \\ \vdots \\ R_{s_m y} \end{bmatrix} R_{yy}^{-1} \tag{54}$$

and the optimal MMSE as a function of b_i is given by equation (55):

$$MMSE : \langle e_k, e_k \rangle = \begin{bmatrix} \tilde{b}_1^* & \tilde{b}_2^* \end{bmatrix} \begin{bmatrix} R_{s_1 s_1} - R_{s_1 y} R_{yy}^{-1} R_{y s_1} & -R_{s_1 y} R_{yy}^{-1} R_{y s_2} \\ -R_{s_2 y} R_{yy}^{-1} R_{y s_1} & R_{s_2 s_2} - R_{s_2 y} R_{yy}^{-1} R_{y s_2} \end{bmatrix} \begin{bmatrix} \tilde{b}_1 \\ \tilde{b}_2 \end{bmatrix} \tag{55}$$

where $R_{s_i y}$ are the cross-correlation matrices of signal y and disturbers s_1 and s_2, R_{yy} is the auto-correlation matrix of signal y, and vectors $\tilde{b}_i$ are vectors formed using filter parameters b_1 and b_2 by adding a variable sample delay.

In order to minimize the error of equation (55), the optimal solution for vectors $\tilde{b}_i$ is found by determining the eigenvector e_{min}, corresponding to the minimum eigen-value of matrix R_c, where R_c is given by equation (56).

$$R_c = \begin{bmatrix} R_{s_1 s_1} - R_{s_1 y} R_{yy}^{-1} R_{y s_1} & -R_{s_1 y} R_{yy}^{-1} R_{y s_2} \\ -R_{s_2 y} R_{yy}^{-1} R_{y s_1} & R_{s_2 s_2} - R_{s_2 y} R_{yy}^{-1} R_{y s_2} \end{bmatrix} \tag{56}$$

Thus, the optimal solution is given by $\left[\begin{array}{c} \bar{b}_1 \\ \bar{b}_2 \end{array}\right]_{opt} = e_{min}$.

Therefore, once the target values for b_1 and b_2 filter outputs are determined by equation (56), then the VEQ w is determined using equation (54). The disturber symbols 1470 are then estimated by the VA 1440.

It should be noted that, in order to reduce the computational complexity of the compensation method, two or more co-channels may be treated as a single channel. Thus, the filters b_1 and b_2 may be treated as one filter by equalizing the two target responses $\bar{b}_1^* = \bar{b}_2^*$ in equation (55) and solving for the VEQ.

In one alternative embodiment of the present invention, each disturber is detected in a successive manner one at a time, as illustrated in Figure 13. In such a successive scheme, it is desirable to design a VEQ that will shorten only one channel while it will suppress the interference from the other disturbers. This can be achieved with the design presented above if only one target response b_1 is used, while nullifying the other target response b_2 such that $b_2 = 0$.

Alternative to VEQ Design

In order to determine the estimates of the disturber symbols $s_i(n)$ of cross-talk interference, a DFE (Decision Feedback Equalizer) method may be used in place of the JVA and the VEQ design. In a DFE, the interference is canceled according to past decisions of the interference symbols. The design of the DFE includes designing two filters: the feedforward filter W and the feedback filter B. It will be appreciated that, in one

-50-

embodiment of the present invention, the W and B filters are designed similarly to the VEQ filter design described above with reference to Figure 14.

Figure 16 illustrates a single user DFE design that may be used in a successive scheme. The disturbers s_1 1601 and s_2 1602 are first modeled by co-channel models h_1 and h_2 1605. The noise 1615 is then added to the summation of the two modeled disturbers 1610. The resulting signal is passed through the W filter 1620. It should be noted that the operation of a DFE is well known in the art and, in order not to obscure the present invention, specific details of the operation of the DFE are not described here. The signal is then passed through a slicer 1630. In order to obtain symbols s_1 1660 of the disturber signal 1601, the output of the slicer 1630 is passed through a feedback filter 1640. The signal is then subtracted from the output of the feedforward filter 1620 and passed through the slicer 1630 in order to obtain the estimated symbols s_1 1660.

It will be appreciated that, in one embodiment of the present invention, the W filter and the B filter may be designed in such a way that a joint DFE may detect the symbols of the summation of the two disturbers s_1 and s_2.

Figure 17 illustrates an alternate embodiment of the joint DFE design where multiple outputs are detected simultaneously. The sum 1710 of the two modeled disturbers s_1 1701 and s_2 1702 is passed through two feedforward filters W_1 1720 and W_2 1725 in order to attempt a first detection of the two disturbers. A multiple input multiple output (MIMO) feedback filter 1770 is designed in order to cancel the past decisions on both the s_1 symbols and s_2 symbols. Thus, at steps 1750 and 1755, the past s_1 and s_2 estimated symbols are subtracted from the output of the feedback filters, resulting in the final estimates symbols s_1

1660 and s_2 1665. The advantage of the MIMO DFE is that it cancels interference according to past decisions and reduces the computational complexity of the compensation method as described herein.

Robust Equalization for Systems with Uncertainty

For cases in which there is dynamic uncertainty in the channel models, the design method described above may be modified to produce an optimal "worst case" equalizer. Generally, the system model includes a nominal dynamics and uncertainty Δ, which are either additive or multiplicative. According to one embodiment of the present invention, an additive form is illustrated in Figure 15, where the uncertainty 1510 is present in the main channel dynamics 1520. The actual channel h_a 1520 can be bounded as: $\|h_a\| \leq \|h\| + \|\Delta\|$, and the uncertainty is generally normalized so that $\|\Delta\| \leq 1$.

Thus, instead of minimizing the mean square error of the estimate, the equalizer 1530 will perform the minimization subject to the bounded uncertainty. In particular, the robust optimal solution will yield a vector pair (w, b) that minimizes the $MMSE : \langle e_k, e_k \rangle$ subject to $\|\Delta\| \leq 1$. It will be appreciated that the various methods described above have robust counterparts which can be solved for using linear matrix inequalities (LMIs), indefinite Riccati solvers, or μ-synthesis techniques.

Performing the compensation method described herein without accounting for uncertainty bounds may lead to compensation with poor performance or system instability. Thus, in order to ensure robustness of compensation and adaptation, uncertainty may be successively reduced at different times of the compensation method described herein. The adaptive compensation method is performed using stochastic plant models and uncertainty

estimated measures acquired with dynamic identification methods. For an example of the computation of uncertainty bounds, see U.S. Patent Application Serial No. 09/345,640, filed June 30, 1999, entitled "Model Error Bounds for Identification of Stochastic Models for Control Design", assigned to the assignee herein, still pending.

Successive Cancellation of Disturbers

Figure 13 illustrates an alternative embodiment of the present invention in which, instead of performing the JVA method, a successive cancellation scheme is used in order to compensate for multiple PAM disturber signals. The successive PAM cancellation method is based on the overall architecture of the compensation system described in Figure 6. Therefore, for a set of disturbers of different rates, block 670 and block 680 of Figure 6 are repeated for each significant disturber signal that has a distinct rate, as illustrated in Figure 13. As such, the DMT signal is first removed from the composite signal 1310 at block 1314. The resulting signal 1315 represents the summation of all PAM disturber signals. For illustrative purposes, Figure 13 shows the compensation of only two PAM signals with sampling rates T1 and T2, however, multiple PAM signals may be used. The signal symbols are estimated separately for the two PAM signals. Additionally, the VEQ design and implementation and the channel tracking is performed separately for the two PAM signals. Once the PAM signals are estimated, they are compensated or subtracted from the composite signal, thus resulting in a compensated DMT signal.

The composite received signal y(k) 1310 is first processed in a standard receiver block 1300. The standard receiver block includes a TEQ 1302, a Prefix/FFT block 1304,

and a FEQ 1306. The operation of these blocks is described above with reference to Figure 3, Figure 4, Figure 5, and Figure 6. The signal is then passed through a slicer function 1312 in order to detect the DMT symbols. Next the signal is remodulated at block 1360 in either the frequency domain, illustrated by blocks 1366 ad 1368, or the time domain, illustrated by blocks 1362 and 1364. In the frequency domain, the operation of the standard receiver block 1300 is reversed by passing the signal through the inverse of the FEQ 1362 and performing an inverse FFT 1364. Similarly, in the time domain, the signal is remodulated using IFFT function 1366 and convolving the channels and the TEQ at block 1368.

In order to obtain a first pass estimation 1315 of the disturber signal, the resulting signal 1370 is then subtracted from the TEQ filtered signal 1309. If the PAM signal rates are not compatible with the DMT received signal, the composite disturber signal 1315 is then resampled at block 1316 at the rate $T_1/2$, for example, of the first PAM signal that will be compensated for. It should be noted that, in one embodiment of the present invention, the successive method described herein detects the high rate PAM disturber first by filtering the low rate PAM disturbers.

At block 1320, a VEQ design on the first PAM disturber signal is performed according to the VEQ design method described above. The PAM symbols are then detected at PAM receiver 1322 and remodulated at block 1324. During data transmission time the actual VEQ implementation and the PAM channel tracking is further performed at block 1326. For an example of PAM channel tracking see co-pending Patent Application Serial No.(), filed on even date herewith, entitled, assigned to the assignee herein.

Next, the resulting estimation of the first PAM signal 1328 is subtracted at block 1330 from the composite disturber signal 1316. The resulting signal is resampled at the rate of the second PAM disturber $T_2/2$. Next, the VEQ design 1330, the estimation of the PAM symbols 1332, remodulation 1334, and tracking 1336 are performed in a similar manner as to the processing of the first PAM signal. Once the second PAM signal is estimated, it is first resampled at the rate of the first PAM disturber, at block 1338, in order to perform a summation of the disturber signals. Finally, the composite estimated disturber signal is processed further by resampling it to the DMT rate at block 1340, prefixing and performing FFT operations at block 1342, and filtering it at block 1344. At block 1350, the estimated disturber signal is subtracted from the composite received signal 1308, thus resulting in the compensated main channel symbols or DMT symbols.

System Transmission/Show Time Phase
1. DMT Removal

As stated earlier, the purpose of the DMT removal block 660 is to extract as much of the DMT component from the received signal as possible, to produce a signal $\hat{y}_{dist}$, which consists mainly of crosstalk disturbance. In one embodiment the DMT Removal block 660 requires two inputs

$$y_{teq}(n) \in R^{544}, \text{ and}$$

$$\hat{q}_1(k) \in C^{256},$$

-55-

that are the received signal after being filtered by the TEQ 620 (in this example, a real vector of 544 points), and the FEQ 640 (in this example, a 256 vector of complex numbers), respectively.

Also as stated earlier, the first step in the DMT removal is an initial (1st Pass) detection on the DMT symbols, as indicated by the slicer function 662 in Figure 6. The 1st pass detection produces symbols, for example in one embodiment 256 symbols, which are then conjugated and modulated back into the time domain using an IFFT (IFFT operation part of block 664). Note, however, that the IFFT results in a real vector of 512 points, 32 points less than a frame of the received vector $y(n)$. This loss of information is due to the cyclic prefix stripping operation that occurs in the standard receiver block. After remodulation, this cyclic prefix is added back to the vector (also part of the operations shown at block 664), and the time vector is then filtered through the equalized main channel $\hat{h}_{teq}$ (also part of the operations shown at block 664). The resulting signal 666 is then subtracted from the received signal to form the 1st pass estimate of the disturbance signal:

$$\hat{y}_{dist}(n) = y_{TEQ}(n) - \hat{y}_{DMT}(n),$$

as shown in Figure 6.

The 1st step in the DMT removal is the first pass detection on the DMT symbols, as indicated by the slicer function 662 in Figure 6. The 1st pass detection is a maximum likelihood symbol-by-symbol slicer. It will be appreciated that the implementation of a slicer is well known to one of ordinary skill in the art and in order to avoid confusion further details of a slicer implementation are not be described herein.

After the 1st pass DMT decisions, the DMT removal module 660 remodulates the 1st pass DMT decisions (part of block 664), filters them through an estimate of the TEQ'd channel (part of block 664), and subtracts off the resulting estimate of the TEQ'd DMT signal 666 from the TEQ output 642.

In order to further illustrate the process of DMT removal, an embodiment of a possible DMT removal subsystem is illustrated in Figure 18. In the embodiment illustrated in Figure 18, the first step in DMT signal removal is conjugate mirroring 1802. DSL modems using DMT use IFFTs to simultaneously modulate QAM symbols onto several subcarriers. Since the output of the IFFT is sent directly to the DAC, then the output of the IFFT must be real-valued. This is accomplished by computing the 512-point IFFT of a complex vector of length 512 in which the upper half contains the 256 QAM symbols to be transmitted and the lower half contains conjugate mirror of these symbols about the Nyquist frequency. In the notation here, $\tilde{q}_1(n)$ shall be the vector of 1st pass DMT decisions and $\tilde{q}_1^m(n)$ shall be the conjugate mirrored vector of decisions (output from block 1802), according to the conjugate mirroring operation as described in a description of the DMT signal above.

After conjugate mirroring 1802 gain and re-modulation is performed on the signal (blocks 1804 and 1806 respectively). After multiplication by a diagonal gain matrix, $\mathbf{G}$, 1804, the output of the IFFT (1806) is

$$\hat{\mathbf{x}}_{dmt}(n) = \mathbf{F}\,\mathbf{G}\,\tilde{\mathbf{q}}_1(n),$$

-57-

where G and the IFFT matrix, F, are both defined as described in a description of the DMT signal above.

After re-modulation is performed a cyclic prefix operation is performed on the signal. Since Fourier transforms are defined for periodic signals, then the received DMT signal must look like a periodic signal for demodulation via an FFT to be accurate. This is accomplished in two steps. First, when the DMT signal is transmitted, it is prefixed with the last N samples of the IFFT output. This prefix is known as the cyclic prefix and in the case of DSL using DMT, $N = 32$. Second, when the DMT signal is received it is passed through a TEQ (time-equalizer) that shortens the impulse response of the channel to 32 samples or less.

Mathematically, the cyclic prefixing step can be expressed as

$$\hat{s}_{dmt}(n) = \Pi \, \hat{x}_{dmt}(n)$$

where, as in the description of the DMT signal above,

$$\Pi = \begin{bmatrix} \underset{32x480}{0} & \underset{32x32}{I} \\ \underset{512x512}{I} \end{bmatrix}.$$

This is thus the time domain estimate of the DMT signal that was transmitted.

The time domain estimate is then reformatted into a serial data stream by a parallel to serial converter 1810. Next, the signal is passed though a filter, which is a model of the DMT channel 1812. This model of the DMT channel 1812 includes any transmit filters

-58-

(e.g., the DAC filter), the twisted-pair line, and any receive filters (e.g., the ADC filter and the TEQ).

An initial estimate of the post-TEQ disturbance signal is then

$$\hat{y}_{dist}(n) = y(n - d_{dmt}) - \hat{y}_{dmt}(n)$$

where y is the TEQ output, and d_{dmt} is an appropriately chosen delay (1814). This delay (1814) exists because the filter $\hat{h}_{teq}(n)$ has a nearly pure delay (the TEQ cursor delay). Note that it may be possible to remove the delay (1814) without a significant loss in re-modulation accuracy provided that the TEQ cursor delay is removed from the TEQ'd channel impulse response, $\hat{h}_{teq}(n)$ 1812. Alternatively, re-modulation may be performed using the FEQ coefficients. Note that when the FEQ coefficients are used to re-modulate the DMT decisions, there is no TEQ cursor delay and hence no signal delay is introduced.

The estimate of the post-TEQ is then refined by the joint Viterbi PAM receiver, which makes use of identified PAM co-channel modules and finite PAM alphabets.

2. Disturber Symbol Detection

As discussed above, the output vector $\hat{y}_{dist}(n)$ 665 from the DMT Removal module 660 is then passed to the Disturber Symbol Detection module 670 where it is run through the VEQ 672 and Joint Viterbi Algorithm 674, resulting in a vector estimate 675 of the disturber symbols $\tilde{s}(n)$.

Following the example given earlier and referring to Figure 6, the VEQ 672 is an FIR

filter with a length, n_w. The output of the VEQ 672 filter is a real vector of the same length

of the input sequence. The processing involved is a convolution of the 544 length input

vector $\hat{y}_{dist}(n)$ 665, and the VEQ w. In one embodiment of the present invention, the length

of the VEQ used is $n_w = 32$ taps.

After processing in the VEQ 672 the ouput vector is processed through the Joint

Viterbi Algorithm (JVA) 674. The Viterbi algorithm employs the shortened co-channels

(which include the effects of the channels and the VEQ) in the computation of the Maximum

Likelihood metric used to search for the optimal disturber input sequence. The design of the

JVA 674 was discussed above, however, it will be appreciated that the implementation of a

JVA is well known to one of ordinary skill in the art and in order to avoid confusion further

details of a JVA implementation are not be described herein.

3. PAM Remodulation and Removal

After the disturber symbol detection block 670 has produced a vector estimate of the

disturber symbols $\bar{s}(n)$, the estimates are routed to the PAM remodulation and removal

block 680. Figure 19 is an illustration of an embodiment of a possible architecture of a

PAM remodulation and removal block, which may also be referred to as a PAM signal

removal system or subsystem. Recall from the discussion above that the purpose of the

PAM signal removal system, such as that shown in Figure 19, is to produce a

compensated version of the 1st pass FEQ output using the PAM decisions produced by

the disturber symbol detection block (Figure 10, block 1040) also called the PAM

receiver. This is accomplished by filtering the PAM decisions through models of the

TEQ'd co-channels (including any pulse-shaping filters, A/D filters, etc.) and summing

the outputs. This time domain estimate of the disturbance signal is then converted to the

frequency domain by computing the FFTs of the appropriate time-windows of data. The

FEQ is applied, as well as any scaling factors such as the AGC gain, and the resulting

compensation vector is subtracted from the appropriately delayed 1st pass FEQ output to

produce the compensated signal. The individual components of the PAM signal removal

system are discussed below.

The joint Viterbi PAM receiver (Figure 6, block 674) produces a sequence of

PAM symbols or decisions for each of the n_c disturbers being compensated. The PAM

co-channel models, however, oversample the PAM baud rate by a factor $P_j, j = 1,..., n_c$.

Hence, it is necessary to upsample the PAM symbols by P_j which introduces zeros

between samples. For example, in illustration Figure 19, $P_j = 4 \ \forall j$, and thus the

upsampled symbol sequences (1902-1 through 1902-C) are

$$\bar{s}_j^/ = (\uparrow 4)\bar{s}_j$$

For the co-channel model filter (1904-1 through 1904-C) an estimate of the

component of the post-TEQ disturbance signal due to the compensated disturbers is

obtained by filtering the upsampled PAM symbol sequences (from blocks 1902-1 through

1902-C) through complete impulse response models of the PAM co-channels, $\hat{h}_{teq,j}$.

Hence, the estimate of the post-TEQ disturbance signal (output from 1906) is

$$\hat{y}_{pam}(n) = \sum_{j=1}^{n_c} \sum_{k=-\infty}^{\infty} \bar{s}_j^f(k) \hat{h}_{ueq,j}(n-k)$$

$$= \sum_{j=1}^{n_c} \sum_{k=-\infty}^{\infty} \bar{s}_j(k) \hat{h}_{ueq,j}(n - P_j k)$$

The complete impulse response model of the PAM co-channel may be obtained during the identification. For an example of identification, see co-pending patent application Serial No. _____, titled "Method and Apparatus for Characterization of Disturbers in Communication Systems", filed on even date herewith and assigned to the assignee herein.

In this embodiment, resamplers Rs (1920-1 through 1920-C) may need to operate on the remodulated signal to convert its data rate to the DMT sampling rate. It should also be noted that all of the above filters may be implemented as FIR filters.

The conversion of the disturber estimate to the frequency domain 1910 is accomplished using the same 512-point FFT that is used to demodulate the DMT signal.

The same FEQ processing that is applied to the received DMT signal (Figure 6, block 640) is applied to the frequency domain representation of the disturber estimate 1912. The FEQ is implemented as 256 complex scalar multiplies.

Next the signal is adjusted for AGC gain 1914, and then, provided the signals have been aligned in time 1916 appropriately as discussed above, the compensated DMT signal (from 1918) with the PAM removed is

$$\hat{q}_2(n) = \hat{q}_1(n) - c(n)$$

-62-

where $\hat{q}_1(n)$ is the original FEQ output.

At this point the compensated DMT signal $\hat{q}_2(n)$, which is in the frequency domain, requires minimal further processing to detect the transmitted symbols and this detection may be performed by a slicer as described above.

4. Showtime Adaptation and Tracking

During showtime the compensation system will enter an adaptive mode which further refines (or updates) co-channel models ($\hat{h}_1$, $\hat{h}_2$), the VEQ, and the shortened channels (b_1, b_2). Adaptation may be desirable when the communications system has changed either because of small changes, for example time drifts and/or thermal/environmental changes, or because of sudden changes, for example a new service turns on or a service turns off. Thus because of these changes it may be desirable to recompute the coefficients of compensation.

For the purposes herein, there are two main types of adaptation: direct adaptation and indirect adaptation. In the present invention direct adaptation is used when there is no explicit tracking of the channels and the design of the compensator does not need to be changed. In other words, only the parameters of the design need to be changed or updated. Thus, in the present invention, direct adaptation adjusts the coefficients of the VEQ and JVA without doing explicit identification or design calculations at showtime. Indirect adaptation is used in the present invention when the channels change such that the design of the compensator needs to be changed. In other words, the parameters of the communication system changed and the design of the compensator must be changed or updated. Thus, in the

present invention, indirect adaptation adjusts the coefficients of the VEQ and JVA by performing identification and design calculations that were previously performed during design time. The extensive computations required for this are executed at a slower rate in real-time.

Figure 20 illustrates a generic block diagram for direct adaptation as applied to the present invention. An adaptation mechanism 2010 takes the input 2055 and output 2056 and some auxiliary signals 2060 and computes and updates the compensator coefficients 2015 and applies those coefficients to the compensator 2050. An example of one embodiment of a direct adaptation mechanism is described in detail below.

Figure 21 illustrates a generic block diagram for indirect adaptation as applied to the present invention. An identification module 2120 takes the input 2155 and output 2156 and some auxiliary signals 2160 and performs an identification process to produce identified model coefficients 2125. Design module 2130 takes the identified model coefficients 2125 and redesigns the compensator coefficients 2135, effectively redesigning the compensator 2150. It should be noted that indirect adaptation is better equipped for the incorporation of uncertainty because the indirect adaptation mechanism includes an identification process.

Direct Adaptation Example

Figure 22 illustrates an exemplary embodiment of a direct adaptation mechanism in the context of the present invention. It should be noted that this example of an adaptation mechanism is merely exemplary and the present invention should not be limited thereto.

In the embodiment illustrated in Figure 22, one or more of the co-channel models ($\hat{h}_1$, $\hat{h}_2$) 2210 and 2211, the VEQ 2220, and/or the shortened channels (b_1, b_2) 2230 and 2231 are

refined (or updated) by using two ongoing least-means-squares (LMS) processes 2250 and 2260. In the embodiment of Figure 22 two distinct LMS routines are envisioned, LMS 2260 for the explicit co-channel models ($\hat{h}_1$, $\hat{h}_2$) for the PAM remodulation, and LMS 2250 for the shortened channels (b_1, b_2) that are required for the Viterbi Algorithm (VA).

In the formulation for the VEQ LMS 2250 the desired objective (cost function) is the expected value of the squared error signal

$$\langle e, e \rangle = E e(n)^2 .$$

The error signal is defined as

$$e(n) = f^T y(n) - \sum_{i=1}^{m} b_i^T s_i(\delta) = \begin{bmatrix} f^T & b_1^T & \cdots & b_m^T \end{bmatrix} \begin{bmatrix} y(n) \\ s_1 \\ \vdots \\ s_m \end{bmatrix} \triangleq g^T u(n)$$

where $f^T \in R^{n_f}$ is the VEQ filter, $b_i \in R^{l_s}$ are the co-channels vectors, s_i are the disturber symbols sent (generally complex), and δ is a relative delay between the shortened channel and the filter path that is used as an optimization degree of freedom.

Normalizing the first tap in order to avoid a trivial solution for the optimal g, it can be shown that the error can be expressed as

$$e(n) = \tilde{g}^T \tilde{u}(n) - s_1(n - \delta);$$

where $\tilde{g}$, $\tilde{u}(n)$ contain vectors b_1, s_1 vectors shortened by one element each.

For real-time implementation s_i are replaced by detected symbols $\hat{s}_i$. In an LMS setting, the gradient of the mean square objective is approximated with the derivative of the instantaneous square of the error (ie, not the averaged value given above...). That is

$$J(n) = |e(n)|^2,$$

and the gradient is taken to be

$$\nabla J(n) = \frac{\partial |e(n)|^2}{\partial \tilde{g}^T} = 2\tilde{u}\tilde{u}^T g - \tilde{u}s_1(n - \delta).$$

It can be shown using algebraic steps that the gradient simplifies to

$$\nabla J(n) = 2\tilde{u}e(n).$$

Thus the update formula for the VEQ based on the LMS becomes:

$$\tilde{g}(n+1) = \tilde{g}(n) - 2\mu\tilde{u}e(n)$$

5. Performance Monitoring

During showtime it may be desirable to monitor the performance of the compensator that was designed so that if the performance drops below a desired level the compensation system may be modified using any of the following techniques: adaptation, retraining, and/or soft reconfiguration.

In one embodiment of the present invention performance monitoring is performed by tracking the errors of the compensated signal. If the errors become large then the compensation system may need to be modified.

In another embodiment of the present invention the errors of the compensated signal are collected and white noise tests are performed on the collected error signals. If the noise of the system is not white (i.e., is "colored"), then there is a disturber (or disturbers) that have not been identified. For examples of white noise tests see Patent Application, Serial Number 09/523,065, titled "Method for Automated System Identification of Linear Systems," filed on March 10, 2000, and assigned to the assignee herein.

In yet another embodiment of the present invention performance monitoring is performed by monitoring the SNR on the main channel in real time. If there are significant changes in the SNR, for example significant changes in the SNR between the first pass of the DMT receiver and the second pass of the DMT receiver, may indicate the compensation design is no longer optimal. Thus, it may be advantageous to re-perform the bit loading calculations of design time because the bit loading structure previously calculated is no longer optimal for the new SNR picture.

6. Retraining

The compensation mechanism of the present invention enters what is referred to as "retraining" when new disturbers appear after the initial training process has occurred. For example, in the DSL system example given above, the compensation system goes through a fast retrain when new services get turned on that interfere with or impair the channels being

compensated. For an example of retraining, see co-pending patent application, Serial Number _____, titled "Method and Apparatus for Cooperative Diagnosis of Impairments and Mitigation of Disturbers in Communication Systems", filed on even date herewith, and assigned to the assignee herein.

7. Soft Reconfiguration

The compute resources available at show time and initialization time are very different compared to a design environment. In the design environment design automation merely assists the designer. The same is not true at show time and initialization time. Real-time design procedures require total automation and very high reliability. To achieve the reliability, the design method can explicitly reflect compute resource costs and constraints in the cost metrics in a real-time planner. Thus, it may be advantageous to have automated design mechanisms (or tools) to perform reconfigurations, for example reconfiguration of the Viterbi algorithm of the present invention during show time, that optimize the design based upon fixed compute costs. For examples of some such reconfiguration tools and methods, see Patent Number 5,880,959, titled "Method for Computer-Aided Design of a Product or Process," issued on March 9, 1999, and assigned to the assignee herein; Patent Application Serial Number 09/345,172, titled "Real-Time Planner for Design," filed June 30, 1999, and assigned to the assignee herein, still pending; Patent Application Serial Number 09/345,166, titled "Adaptation to Unmeasured Variables," filed June 30, 1999, and assigned to the assignee herein, still pending; and Patent Application Serial Number 09/345,640, titled

"Model Error Bounds for Identification of Stochastic Models for Control Design," filed June 30, 1999, and assigned to the assignee herein, still pending.

8. Fixed Point Implementation

In some embodiments of the present invention it may be desirable and/or necessary to design a tight footprint implementation. For example, in the DSL system example given above, if the compute resources and/or hardware resources are limited on the transceiver it may be necessary to decrease the compute costs and/or footprint of the compensation software and architecture in an automated design procedure. One way to reduce the footprint and/or compute cost of the present invention is to implement the compensator in a fixed point implementation. For an example of a method and apparatus for fixed point implementation see co-pending patent application Serial Number _____, titled "Method and System for Design and Implementation of Fixed-Point Filters for Control and Signal Processing," filed on even date herewith, and assigned to the assignee herein.

Thus, a method and system for compensating for cross-talk interference in communication systems have been described. Although specific embodiments have been described, various modifications to the disclosed embodiments will be apparent to one of ordinary skill in the art upon reading this disclosure. Therefore, it is to be understood that the specification and drawings are merely illustrative of and not restrictive on the broad invention and that this invention is not limited to the specific embodiments shown and described.

CLAIMS

What is claimed is:

1. A method comprising:

   determining an estimation of at least one interfering signal; and

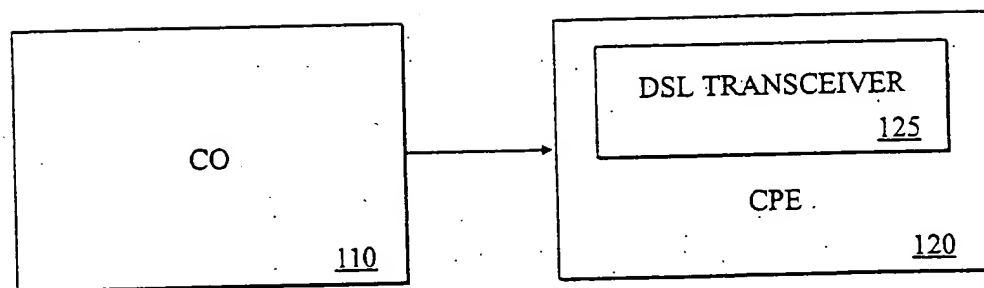   performing a compensation operation on the at least one interfering.
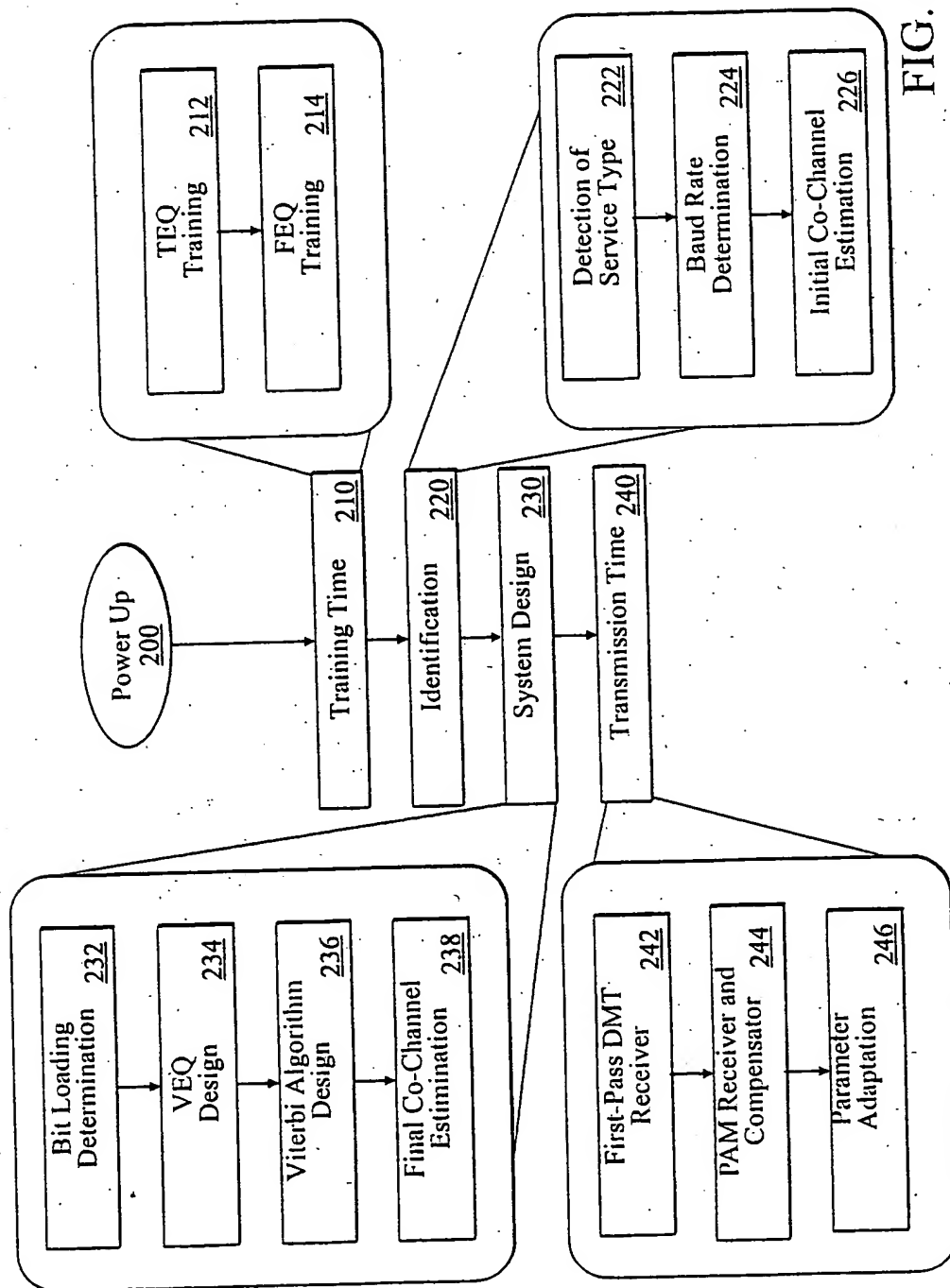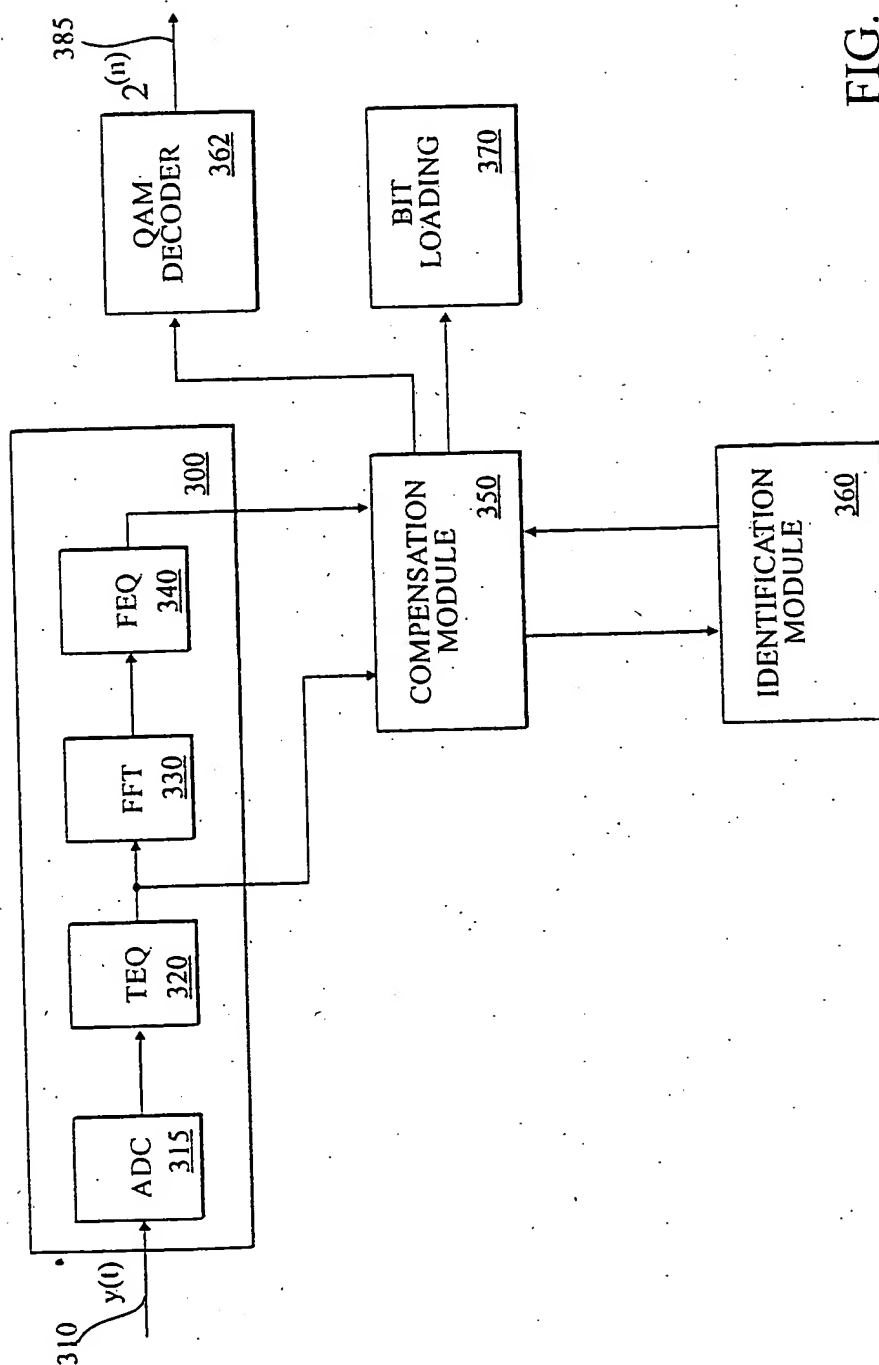
1/22

FIG. 1

2/22

FIG. 2

Power Up
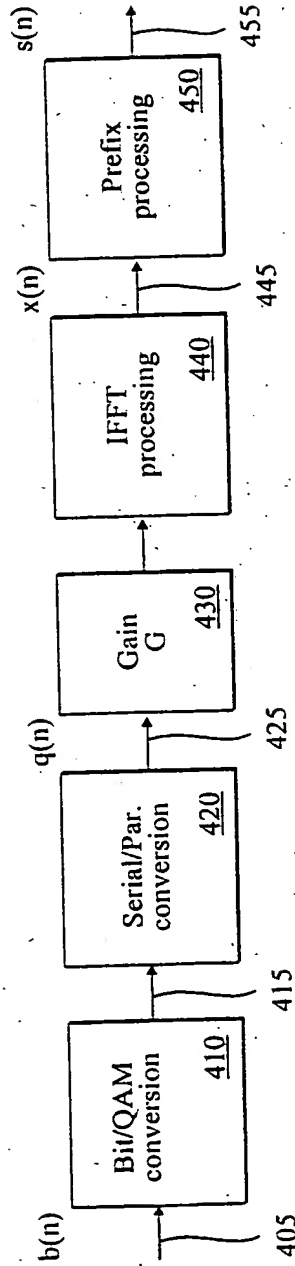200

Training Time 210

Identification 220

System Design 230

Transmission Time 240

TEQ Training 212

FEQ Training 214

Detection of Service Type 222

Baud Rate Determination 224

Initial Co-Channel Estimation 226

Bit Loading Determination 232

VEQ Design 234

Viterbi Algorithm Design 236

Final Co-Channel Estimination 238

First-Pass DMT Receiver 242

PAM Receiver and Compensator 244

Parameter Adaptation 246

3/22

FIG. 3

4/22

FIG. 4

FIG. 5

5/22

FIG. 6

6/22

FIG. 7

7/22

Use $S_2 = S_1$ in equation (22), and evaluate equations (22) - (26) to initialize the PAM SNR S_p.

<u>810</u>

Use the current value of S_p to evaluate equations (27)-(29) and use the resulting value of S_2 to evaluate equations (22)-(26).

<u>820</u>

yes

Is S_p is larger than a selected lower bound?

<u>830</u>

NO

The predicted compensated SNR S_2 is the one that corresponds to the value of S_p.

<u>840</u>

FIG. 8

8/22

$S_i, \Gamma, D_i^{PAM}, D_i^{tot}$

↓ 900

┌─────────────────────────┐
│ Disturber Ranking   │
│   │
│   910   │
└─────────────────────────┘

S_i, Γ_v

↓ 930

┌─────────────────────────┐   ┌─────────────────────────┐
│ Predicted Bit Loading (b_p) │   │ Viterbi-limited Bit Loading │
│   │   │   (b_v)   │
│   920   │   │   940   │
└─────────────────────────┘   └─────────────────────────┘

942

┌─────────────────────────┐
│ Select Final Bit Loading (b_f) │ ──→ Final Bit
│   │   Loading (b_f)
│   950   │
└─────────────────────────┘

↓

┌─────────────────────────┐
│ Select Final Gains (g_f) │ ──→ Final Gains (g_f)
│   │
│   960   │
└─────────────────────────┘

FIG. 9

9/22

```
┌─────────────────────────┐
│                         │
│  INPUTS                 │
│                         │
│  • TOTAL CONSTELLATION  │
│  • SET OF CO-CHANNELS    │─────┐
│  • DES. RED CONSTELLATION SIZE │
│                   1010  │     │
└─────────────────────────┘     │
                                 ▼
                    ┌─────────────────────────┐
                    │                         │
                    │     AGGREGATION         │──────┐
                    │                         │      │
                    │                  1020   │      │
                    └─────────────────────────┘      │
                                                      ▼
                                         ┌─────────────────────────┐
                                         │                         │
                                         │   AGGREGATED            │
                                         │   CONSTELLATION         │
                                         │                         │
                                         │                  1030   │
                                         └─────────────────────────┘
```
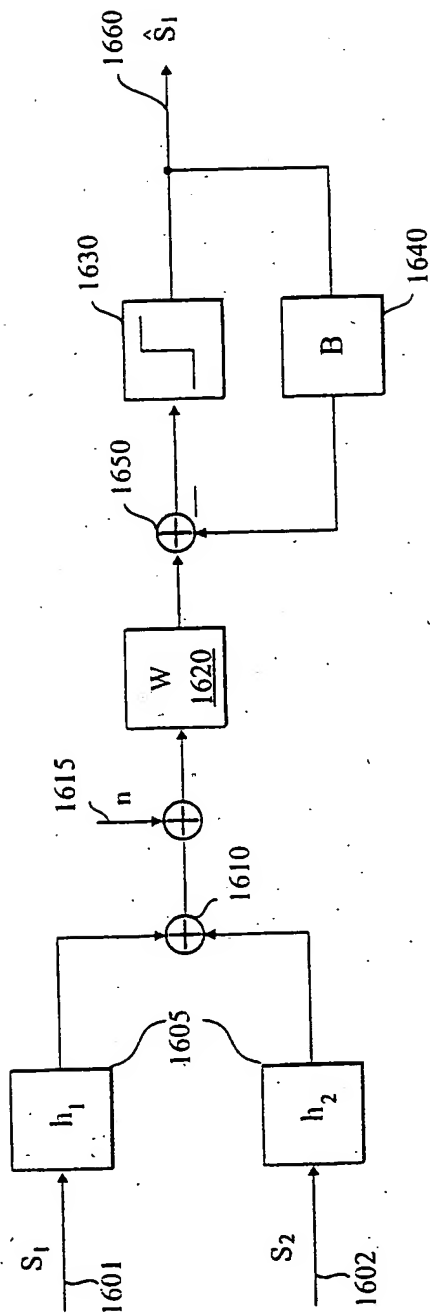
FIG. 10
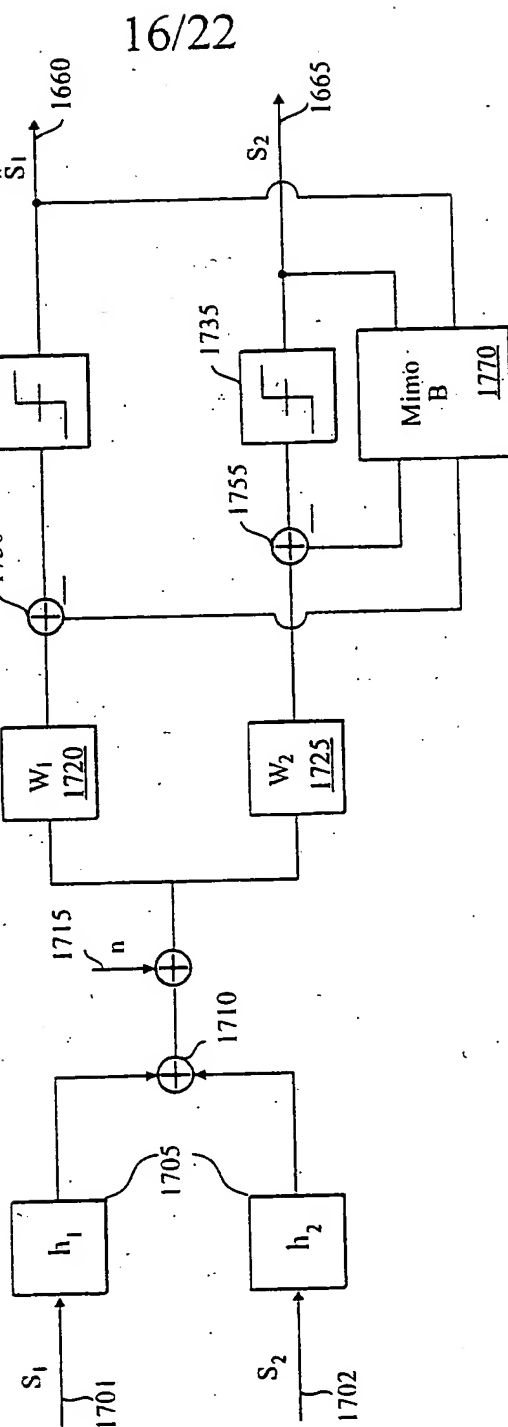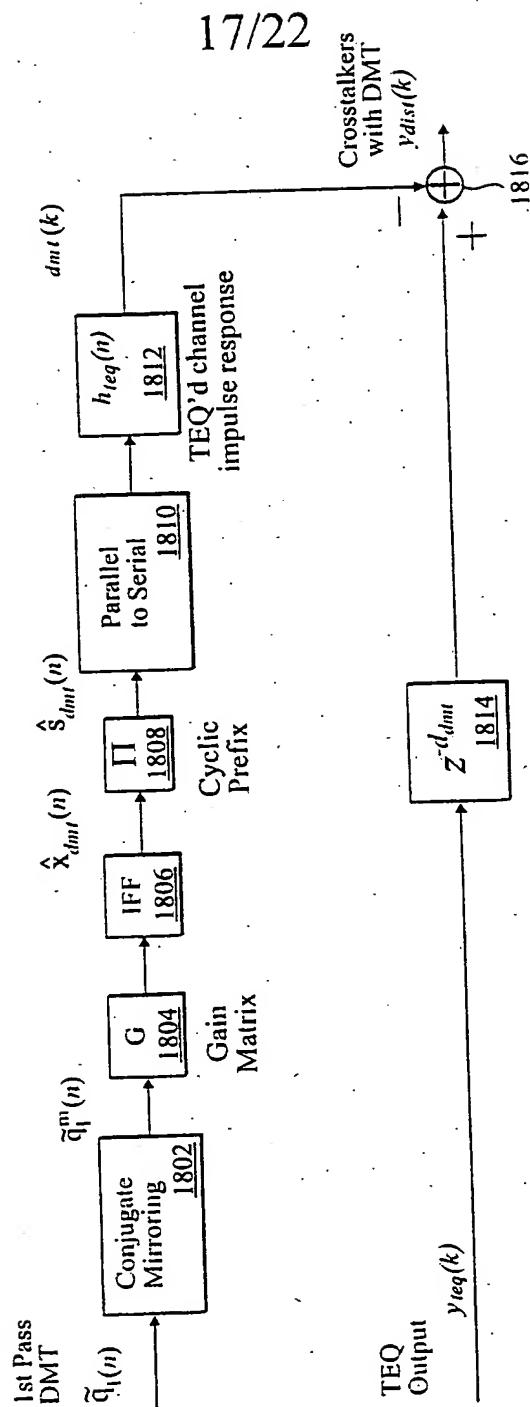
FIG. 11

11/22

FIG.12

FIG. 13

FIG.14

FIG. 15

15/22

FIG. 16

16/22

FIG.17

17/22

FIG. 18

18/22

FIG. 19

FIG. 21

INDIRECT ADAPTATION

Compensator 2150

2155

2156

Compensator Coefficients 2135

Design Module 2130

Identified Model Coefficients 2125

Identification Module 2120

Aux Signals 2160

FIG. 20

DIRECT ADAPTATION

Compensator 2050

2055

2056

Compensator Coefficients 2015

Adaptation Mechanism 2010

Aux Signals 2060
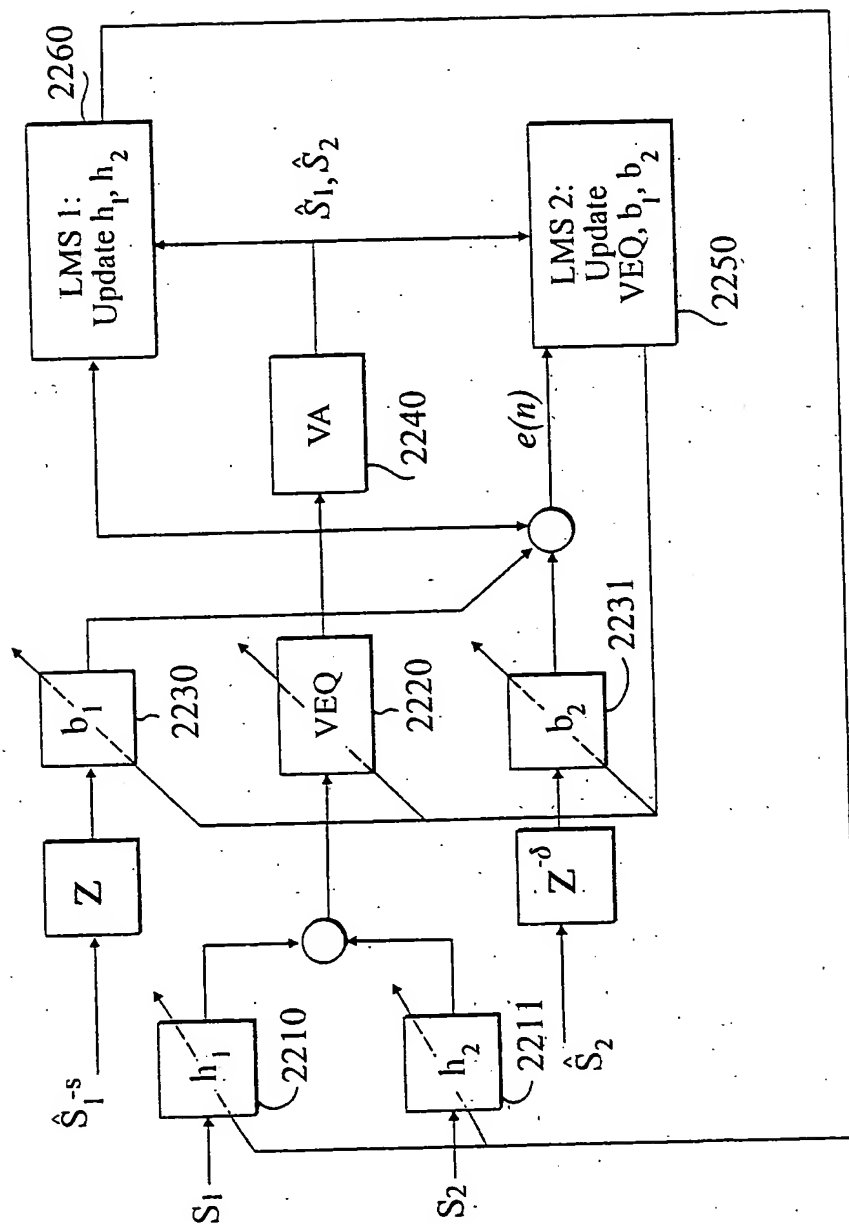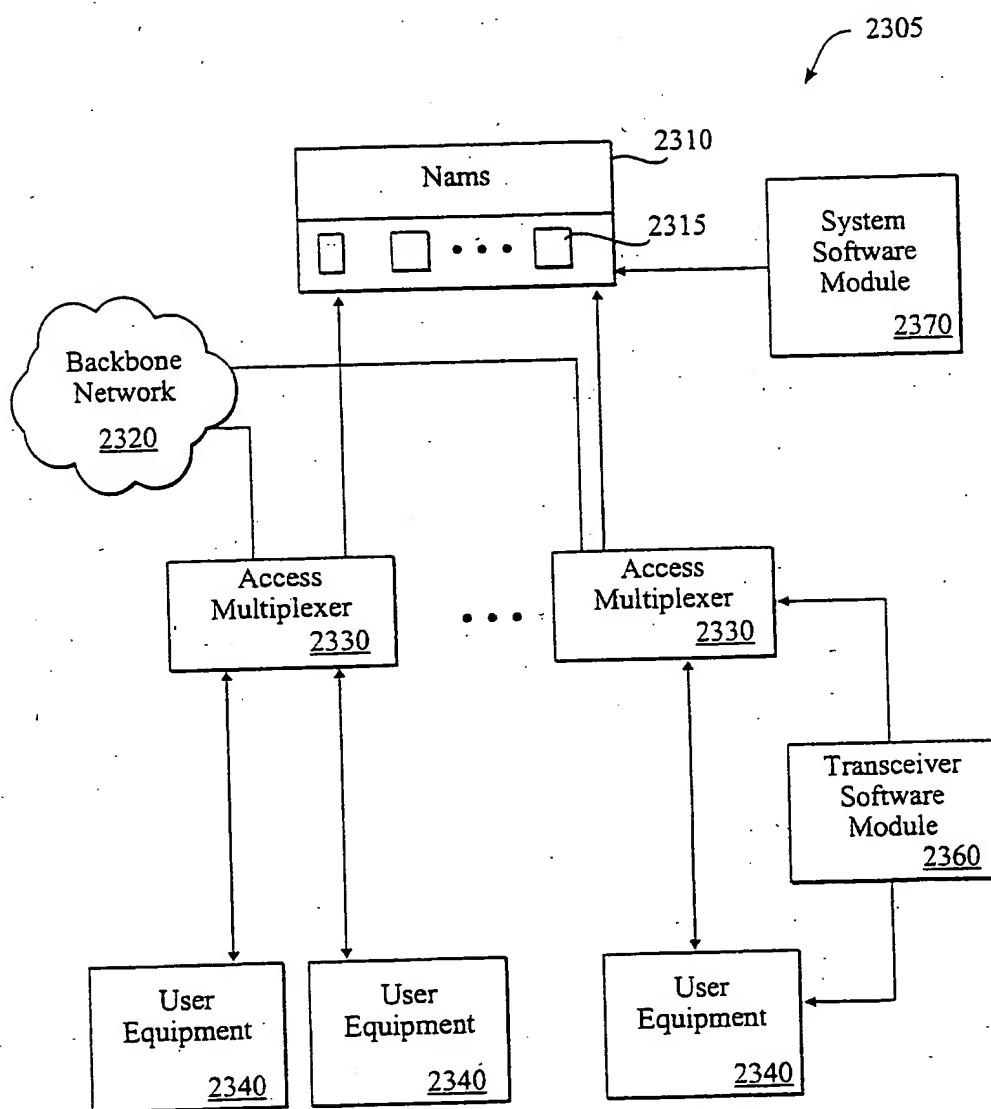
FIG. 22

FIG. 23

22/22

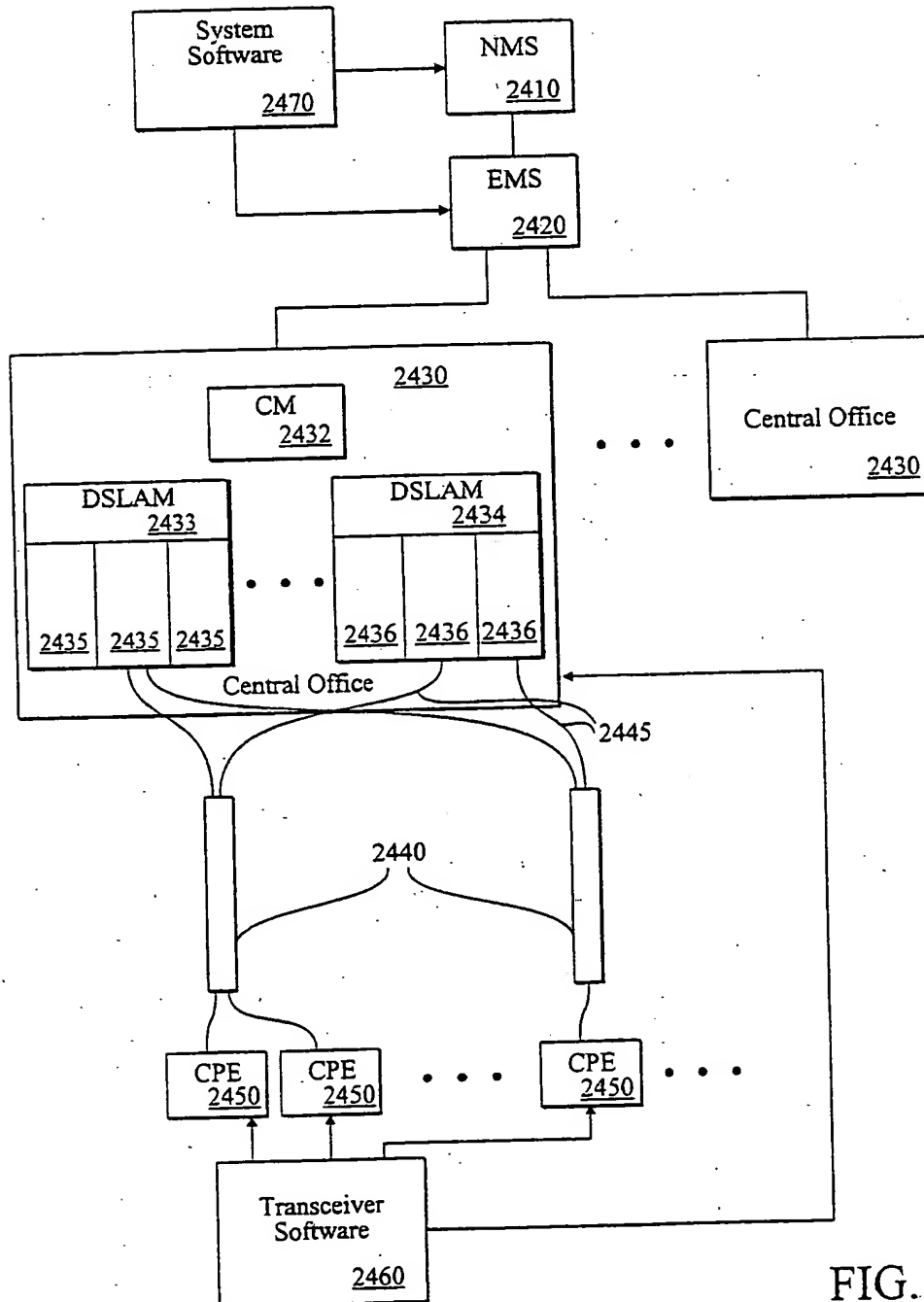FIG. 24

International application No.

PCT/US00/30859

A. CLASSIFICATION OF SUBJECT MATTER

IPC(7) : H04M 1/24

US CL :379/1

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 379/1, 10, 113, 220, 414, 416, 417

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

EAST

SEARCH TERMS: INTERFERENCE OR CROSSTALK, ESTIMATION, COMPENSATION, DIGITAL SUBSCRIBER LINE

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 5,887,032 A (CIOFFI) 23 March 1999, Abstract, column 12, line 16-column 14, line 49.	1
X, P	US 5,995,566 A (RICKARD et al) 30 November 1999, Abstract, column 3, line 63-column 4, line 19.	1
A, P	US 6,055,297 A (TERRY) 25 April 2000, Abstract, column 5, line 56-column 7, line 25.	1

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be of particular relevance	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"E"	earlier document published on or after the international filing date		
"L"	document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)	"Y"	document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"O"	document referring to an oral disclosure, use, exhibition or other means		
"P"	document published prior to the international filing date but later than the priority date claimed	"&"	document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
31 DECEMBER 2000	**27 MAR 2001**

Name and mailing address of the ISA/US	Authorized officer
Commissioner of Patents and Trademarks	AHMAD MATAR
Box PCT	
Washington, D.C. 20231	
Facsimile No. (703) 305-3230	Telephone No. (703) 305-4731

Form PCT/ISA/210 (second sheet) (July 1998)*